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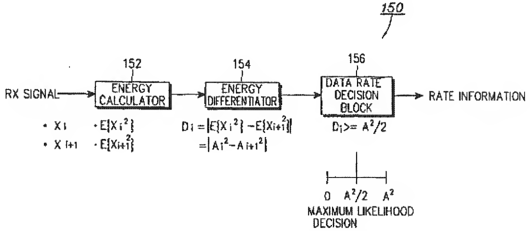
(54) Title: DATA RATE DETECTION DEVICE AND METHOD FOR A MOBILE COMMUNICATION SYSTEM



(57) Abstract: A data rate detecting device detects a data rate for a received signal based on a variation of the energy for the respective received signals between the two adjacent intervals upon failure to receive information about the data rate, and performs channel decoding of the detected data rate information. First, the data rate detecting device divides an interval defined as between a lowest and highest one of a plurality of given data rates into m discriminating intervals. Then, the device calculates a difference between an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer is less than m. If the difference between the average energies is greater than or equal to a threshold value, the device determines that the received signal in the (i+1)'th discriminating interval is transmitted at a data rate corresponding to the i'th discriminating interval.

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DATA RATE DETECTION DEVICE AND METHOD FOR A MOBILE COMMUNICATION SYSTEM

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates generally to a channel signal receiving device and method for a mobile communication system, and more particularly, to a device and method for detecting the data rate of a received signal.

2. Description of the Related Art

Code division multiple access (CDMA) mobile communication systems have developed from the conventional mobile communication standard, which focused on voice service, to the IMT-2000 standard, which provides high-speed data transmission. The IMT-2000 standard encompasses various services, including high quality voice, moving pictures, and Internet browsing. Communication links provided between a mobile station and a base station in the CDMA mobile communication system are generally classified into a downlink (DL), directing data to the mobile station from the base station, and an uplink (UL), directing data to the base station from the mobile station.

For voice or data transmission on the downlink or uplink, the data rate of the data may dynamically vary periodically, where the period is a predetermined time, e.g., 10 msec, which depends on the type of service. Usually, information about the data rate is transmitted to a receiver and used for decoding. However, in the event the receiver fails to receive the information about the data rate, the receiver has to detect rate of the received signal actually transmitted from the transmitter by analyzing the received signal. This procedure, where the receiver detects the data rate from the received signal, is called "blind rate detection (BRD)".

A description is provided herein below for a BRD operation according to the prior art which is performed in the case of voice transmission using convolutional codes for the purpose of forward error correction (FEC).

First, it is assumed that a set of data rates of voice data which a receiver (i.e., mobile station) uses to service a transmitter (i.e., base station) is designated as $R = \{R_1, R_2, \ldots, R_n\}$, where the data rates are listed in ascending order. To detect an actual data rate R_a reported by the transmitter, the receiver performs a Viterbi decoding of the data from the lowest data rate R_1 and then checks cyclic redundancy codes (CRC's). If the result of CRC check for R_1 is "good", there is a high probability that $R_a = R_1$, and R_a is determined as the actually transmitted data rate to be R_1. If the result of the CRC check for R_a is "bad", the receiver continues a Viterbi decoding of additional data up to the next data rate R_2, i.e., at a data rate $(R_2 - R_1)$, followed by CRC checks. As an attempt to reduce a false alarm potential of the BRD operation, the receiver checks an internal metric for Viterbi decoding, in addition to the CRC check.

As described above, the receiver first performs a Viterbi decoding and then a CRC check in order to detect a rate of convolution coded voice data. The BRD operation, however, is not easy to apply in the case of data transmission using turbo codes. This is because, unlike the Viterbi decoder, a turbo decoder has an internal turbo de-interleaver the type of which is dependent on the data rate. Specifically, when the result of CRC check at a given data rate is "bad", the turbo decoder has to repeat the data decoding process from the first data rate in order to check the CRC for a next data rate, while the Viterbi decoder has only to read additional data to the next data rate and then continue the data decoding. Another reason why the BRD operation is inadequate to the turbo decoder is in that the turbo decoding is usually performed iteratively, with the maximum number of iterations for a data rate being about 8 to 12, which leads to an increase in complexity of the decoder and which takes a long delay time when the iterative decoding is performed for CRC checks at all data rates.

SUMMARY OF THE INVENTION

It is, therefore, an object of the present invention to provide a device and method for detecting a data rate from a received signal upon failure to receive information about the data rate in a mobile communication system.

It is another object of the present invention to provide a device and method for detecting a data rate upon failure to receive information about the rate of turbo coded data.

It is yet another object of the present invention to provide a device and method for detecting a data rate not received during transmission of convolutional coded or turbo coded data.

It is still another object of the present invention to provide a device and method for reducing complexity of a data rate detecting operation upon failure to receive information about the data rate.

To achieve the above objects of the present invention, a data rate detecting device is provided that detects a data rate for a received signal based on a variation of the energy for the respective received signals between the two adjacent intervals upon failure to receive information about the data rate, and performs channel decoding of the detected data rate information.

The data rate detecting device first divides an interval defined as between a lowest and highest one of a plurality of given data rates into m discriminating intervals. Then, the device calculates a difference between an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer and is less than m. If the difference between the average energies is greater than or equal to a threshold, the device determines that the received signal in the (i+1)'th discriminating interval is transmitted at a data rate corresponding to the i'th discriminating interval.

BRIEF DESCRIPTION OF THE DRAWINGS

The above and other objects, features and advantages of the present invention will become more apparent from the following detailed description when taken in conjunction with the accompanying drawings in which:

Fig. 1 is a schematic block diagram illustrating a decoder for a mobile communication system including a data rate detector in accordance with the present invention;

Fig. 2 is a diagram illustrating a data rate detecting operation in accordance with the present invention;

Fig. 3 is a detailed block diagram illustrating the data rate detector shown in Fig. 1;

Fig. 4 is a flow chart illustrating the (i+1)'th interval data rate detecting operation in accordance with the present invention; and

Fig. 5 is a flow chart illustrating the i'th interval data rate detecting operation in accordance with the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Turning to the drawings, a preferred embodiment of the present invention is described hereinbelow in detail with reference to the accompanying drawings. In the following description, well-known functions or constructions are not described in detail to avoid obscuring the invention in unnecessary detail.

Fig. 1 is a schematic block diagram of a decoder of a mobile station receiver in a mobile communication system including a data rate detector in accordance with the present invention. The invention is applicable to any CDMA mobile communication system, such as universal mobile telecommunication system (UMTS), CDMA2000, etc.

Referring to Fig. 1, a de-interleaver 110 de-interleaves a received signal to generate a de-interleaved signal (symbol) X_k. Discontinuous transmission (DTX) bit extractor 120 extracts, from the de-interleaved signal X_k, bits indicating a discontinuous transmission mode received from a base station in a discontinuous transmission mode of the mobile communication system. Data rate detector 150 detects a variable data rate of the received signal (symbol) X_k de-interleaved at the de-interleaver 110, ultimately detecting the rate of the received data upon the failure to receive information regarding the data rate. Specifically, the data rate detector 150 measures variations of energy for each received signal in two adjacent intervals and detects the data rate of the received signal based on the result of detection. The information about the data rate detected at the data rate detector 150 is applied to a rate matching block 130 and a channel decoder 140. The rate matching block 130 receives the de-interleaved symbols to perform a reverse process of puncturing, i.e. symbol insertion, and a reverse process of repetition, i.e. symbol combining, thus generating rate-matched symbols. Channel decoder 140 decodes the rate-matched symbols received from the rate matching block 130. The channel decoder 140 may be implemented with a convolutional decoder or a turbo decoder. The rate matching block 130 and the channel decoder 140 use the data

rate information received from the data rate detector 150 to perform rate matching and channel decoding operations.

Fig. 2 is an illustration for explaining a data rate detecting operation of the present invention performed at the data rate detector 150 shown in Fig. 1.

First, it is assumed that the number of symbols received at the mobile station receiver varies in the order of R_1, R_2, R_3, R_4 and R_5 over time, as shown in Fig. 2. A change in the number of symbols by the interval, e.g., 10 msec means that the data rate is variable. Thus it should be noted that the term "the number of symbols" is substantially interchangeable with the term "data rate".

Fig. 2 shows a case where the base station transmitter correctly transmits data for intervals 1 to 4 but fails to transmit data between intervals 4 and 5. The data symbols in the transmission intervals 1 to 4 are de-interleaved at the de-interleaver 110 shown in Fig. 1 and stored in an internal buffer of the DTX(Discontinuous Transmission) bit extractor 120. Between the intervals 4 and 5, the base station transmitter sends DTX bits in a DTX mode. For such a DTX interval, the base station transmitter disables the transmission power and only an additive white Gaussian noise (AWGN) exists. So, the data rate is R_4 for the DTX interval 5. As such, the present invention uses a fundamental principle that involves determination of a presence of the data in substantially non-transmission intervals for data or data rate information, and ultimately detection of the data rate.

Now, a detailed description will be given to the principle of the data rate detection according to the present invention.

Expediently, it is assumed that there are two data rates R_1 and R_2. In such a case, the following equations may be used in order to determine, without receiving any data rate information, whether a signal has been transmitted at R_1 or R_2. When the received signal from bit position 1 to bit position R_1 is X_1, and the received signal from bit position (R_1+1) to bit position R_2 is X_2, the signals X_1 and X_2 are expressed by:

[Equation 1]

$$X_1 = A_1 \times a_1 + n_1$$

$$X_2 = A_2 \times a_2 + n_2$$

In Equation 1, A_1 and A_2 represent transmission power levels of the signals transferred from the base station transmitter to the mobile station receiver and correspond to $\pm A$ in the presence of the signals or "0" for DTX; a_1 and a_2 represent Rayleigh random variables having a probability function of $p(a_1) = 2 \times a_1 \times \exp(-a_1^2)$ or $p(a_2) = 2 \times a_2 \times \exp(-a_2^2)$, respectively; and n_1 and n_2 represent AWGN random variables with mean "0" and variance σ^2. If the noise variance of the transmission channel is σ^2, the interval-based energy (power) of the received signal is given by:

[Equation 2]

$$E\{X_1^2\} = A_1^2 + \sigma^2$$

$$E\{X_2^2\} = A_2^2 + \sigma^2$$

The differentiation equation of the energies $E\{X_1^2\}$ and $E\{X_2^2\}$ of the received signals gives D_1 as expressed by:

[Equation 3]

$$D_1 = \left|E\{X_1^2\} - E\{X_2^2\}\right| = \left|A_1^2 - A_2^2\right|$$

In Equation 3, if $A_1^2 = A_2^2$, D_1 is "0"; otherwise, if $A_2^2 = 0$ (i.e., for DTX), D_1 is "A_1^2". Namely, when the actual data rate is R_2, D_1 nearly reaches "0"; otherwise, when the actual data rate is R_1, D_1 becomes almost "A_1^2".

The above equations can be applied only if the secondary probability characteristic, average deviation σ^2 is known irrespective of the probability functions $p(a_1)$ and $p(a_2)$ of the Rayleigh random variables. It is of cause assumed that the random variables is not time varying. For reference, the differentiation result of the energies of the received signals $D_1 = \left|E\{X_1^2\} - E\{X_2^2\}\right|$ can be calculated from a given interval-based energy of the individual received signals. The most important variable in determining D_1 may be the accumulation of data sufficient to determine the average energy value. An accurate data rate may be determined when the minimum data rate R_1 is 32 kbps, i.e., the data transmitted in the 10msec frame interval is more than 320 bits.

The above-stated data rate detecting operation can be generalized as follows.

First, it is assumed that a set of serviceable data rates is designated as $R = \{R_1, R_2, \ldots, R_n\}$, in which the data rates are listed in the ascending order. Information about the serviceable data rates is called "transport format set (TFS)" given to the mobile station by the base station in a call setup phase. If information about n data rates is given, one interval is first assigned to the largest data rate R_n and (n-1) intervals are assigned to the other data rates. To be differentiated from the interval assigned to the largest data rate R_n, (n-1) intervals for the other data rates are defined as discriminating intervals. The data rate of the received signal for the individual is detectable. For instance, an average energy of the received signals up to the i'th discriminating interval is subtracted from an average energy of the received signals up to the (i+1)'th discriminating interval. The resulting subtracted value is compared to a predetermined threshold to detect the data rate of the received signal for the (i+1)'th interval.

Now, the operation of detecting the data rate of the received signal for the (i+1)'th interval is described in connection with generalized expressions as follows. A received signal up to the i'th interval designated as X_i can be defined as:

[Equation 4]

$$X_i = A_i \times a_i + n_i$$

In Equation 4, A_i represents the transmission power level of the base station transmitter and correspond to $\pm A$ in the presence of the signal or "0" for DTX; and a_i and n_i represent the Rayleigh random variable and the AWGN random variable as defined above, respectively. From Equation 3, a decision criterion can be defined as in Equation 5 below, from 1 to n. When the received signal up to the i'th interval is X_i and a received signal up to the (i+1)'th interval is X_{i+1}, the differentiation result of the energies $E\{X_i^2\}$ and $E\{X_{i+1}^2\}$ of the received signals gives D_i as expressed by:

[Equation 5]

$$D_i = \left|E\{X_i^2\} - E\{X_{i+1}^2\}\right| = \left|A_i^2 - A_{i+1}^2\right|$$

In Equation 5, if the data are continuously transmitted up to the (i+1)'th interval, i.e., $A_i^2 = A_{i+1}^2$, then D_i is "0"; otherwise, if the data are transmitted up to the i'th interval but not transmitted from the i'th to the (i+1)'th interval (for DTX), i.e., $A_{i+1}^2 = 0$, then D_i is "A_i^2". Therefore, during DTX ($A_{i+1}^2 = 0$), the mobile

station receiver searches for the initial index i and considers the corresponding R_i as the actual data rate for the received data from the base station transmitter.

Fig. 3 is a schematic block diagram of the data rate detector 150 shown in Fig. 1, in which the data rate detector 150 comprises an energy calculator 152, an energy differentiator 154 and a data rate decision block 156.

Referring to Fig. 3, the energy calculator 152 measures energy E_i for a received signal X_i up to the i'th interval and energy E_{i+1} for a received signal X_{i+1} from the i'th interval to the (i+1)'th interval. Namely, the energy calculator 152 accumulates the received signals up to the i'th interval and the received signals up to the (i+1)'th interval to calculate energies E_i and E_{i+1} for the respective received signals X_i and X_{i+1} according to Equation 6 below, which is used to calculate energy E_{i+1} for the received signal X_{i+1}.

[Equation 6]

$$E_{i+1} = \frac{1}{R_{i+1} - R_i} \sum_{k=R_i}^{R_{i+1}} X_k^2 dk$$

The energy differentiator 154 calculates a difference (D_i) between energy $E\{X_i^2\}$ in the i'th interval and energy $E\{X_{i+1}^2\}$ in the (i+1)'th interval, as obtained in Equation 6. The difference between the energies $E\{X_i^2\}$ and $E\{X_{i+1}^2\}$ may be expressed as a difference between the squares of the transmission power levels, as defined in Equations 3 and 5, i.e., a difference between a square A_i^2 of the transmission power level of a received signal for the i'th interval in the i'th interval, and a square A_{i+1}^2 of the transmission power level of a received signal for the (i+1)'th interval. The data rate decision block 156 determines the rate of the transmission data using the energy difference D_i calculated at the energy differentiator 154. If D_i is a desired value A_i^2 as in Equation 5, the data rate decision block 156 determines the data rate R_i for the i'th interval as the rate of the presently transmitted data.

However, considering the actual channel environment, it is impossible that the energy difference between the two intervals as designated by D_i is "0" or A_i^2. That is, the difference D_i itself is a probability variable, where the conditional expectation of D_i satisfies $E\{D_i | A_i^2 = A_{i+1}^2\} = 0$ and $E\{D_i | A_i^2 \neq A_{i+1}^2\} = A^2$. Thus, the data rate decision block 156 compares the energy difference D_i between the two adjacent intervals with a threshold value to determine the data rate. More particularly, the data rate decision block156 determines the data rate R_i for the

previous interval, the i'th interval as the data rate for the current interval when the energy difference D_i between the two adjacent intervals is less than or equal to the threshold value. The threshold value can be designated as a medium value between "0" and A^2, i.e., $A^2/2$ according to a maximum likelihood (ML) principle. Here, A denotes the transmission power level of the received signal from the base station transmitter and $A^2/2$ is half the transmission power level of the received signal. The information about the data rate determined by the data rate decision block 156 is applied to the rate matching block 130 and the channel decoder 140, as shown in Fig.1.

The flow chart of Figs. 4 and 5 illustrate a data rate detecting operation using the above equations which is performed at the data rate detector 150 shown in Fig. 3. Fig. 4 is a flow chart illustrating an operation of detecting the data rate for the (i+1)'th interval from the energies of the received signals for the two adjacent intervals, the i'th and (i+1)'th intervals. Fig. 5 is a flow chart illustrating a general operation of detecting the data rate for the i'th interval.

Referring to Fig. 4, the data rate detector 150 shown in Fig. 1 calculates the energy difference D_i between the two adjacent intervals for each iteration and compares the energy difference D_i with a threshold value $A^2/2$. The data rate detector150 estimates the data rate R_i for the i'th interval as an actual data rate R_{est}, in step 405, when the energy difference D_i is greater than or equal to the threshold value.

Specifically, the energy calculator 152 shown in Fig. 3 accumulates received signal X_i between the (i-1)'th interval and the i'th interval in step 401, and calculates energy $E\{X_i^2\}$ for the received signal X_i in step 402. Also, the energy calculator 152 accumulates received signal X_{i+1} between the i'th interval and the (i+1)'th interval and calculates energy $E\{X_{i+1}^2\}$ for the received signal X_{i+1} in step 402. The energy differentiator 154 calculates an energy difference between the two adjacent intervals, in step 403. That is, the energy differentiator 154 determines the energy difference between the two intervals as $D_i = \left|E\{X_i^2\} - E\{X_{i+1}^2\}\right|$. As previously stated, the energy difference can also be expressed as $D_i = \left|A_i^2 - A_{i+1}^2\right|$. In step 404, the data rate decision block 156 compares the energy difference between the two adjacent intervals with a threshold value, i.e., it determines whether the energy difference D_i is greater than or equal to the threshold value $A^2/2$. When the energy difference D_i is greater than or equal to the threshold value $A^2/2$, the data rate decision block 156 estimates the data rate R_i for the i'th interval

as the actual data rate R_{est} for the current (i+1)'th interval, in step 405. The estimated data rate is provided to the DTX bit extractor 120, the rate matching block 130 and the channel decoder 140, as shown in Fig. 1, and used for rate matching and channel decoding operations.

Referring to Fig. 5, in step 501, the data rate detector initializes the search interval i to "1" and the average power (energy) for the previous interval $E\{X_{i-1}^2\}$ to "0". The energy calculator 152 shown in Fig. 3 calculates, in step 502, the average power for the search interval 1, i.e., first calculates the average power for the current interval $E\{X_i^2\}$. In step 503, the energy differentiator 154 calculates (a second calculation) an energy difference between the previous interval and the current interval according to discriminating equation D_{i-1}. If the data rate decision block 156 determines in step 504 that the result of discriminating equation D_{i-1} is greater than or equal to the threshold value $A^2/2$ (where, the data rate means "0" kbps as i = 1), the data rate decision block 156 estimates the data rate for the current interval R_{est} as the data rate for the previous interval (R_{i-1}) in step 508.

Otherwise, i.e., if it is determined in step 504 that the result of discriminating equation D is less than the threshold value $A^2/2$, the data rate decision block 156 stores the average power $E\{X_i^2\}$ for the current interval in the average power $E\{X_{i-1}^2\}$ for the previous interval in step 505, and increases i by one for searching the next interval in step 506. The energy calculator 162 in step 507 calculates (a third calculation) the average power in the interval i+1 and then stores the calculated average power in the average power $E\{X_i^2\}$ for the current interval. The process returns to step 503 to calculate the discriminating equation D_{i-1} based on the average power $E\{X_i^2\}$ and compares in step 504 the result value of the discriminating equation D_{i-1} with the threshold value.

While repeating the above procedures, when it is determined as $D \geq A^2/2$ in step 504, the data rate decision block 156 estimates the data rate R_{est} of the current interval as the data rate R_{i-1} up to the previous interval.

As described above, the present invention estimates a data rate for a received signal prior to a decoding operation even when no information about the data rate is received from the base station transmitter, which reduces the complexity as compared to the conventional BRD operation which detects the data rate after Viterbi decoding and the CRC check. The present invention thereby reduces the complexity in detecting the rate of turbo-encoded data without a need of a rate-based decoding operation, in the worse case, as often as the maximum number of iterations.

Furthermore, the present invention determines the data rate using only accumulated statistics, irrespective of the type of the channel encoder, and is thus compatible with any channel encoder. For example, even with a convolutional encoder is used, the present invention makes it possible to estimate the data rate with reliability for a frame whose data rate is not less than a threshold value.

While the invention has been shown and described with reference to a certain preferred embodiment thereof, it will be understood by those skilled in the art that various changes in form and details may be made therein without departing from the spirit and scope of the invention as defined by the appended claims.

WHAT IS CLAIMED IS:

1. A method for detecting a data rate in a mobile communication system, comprising the steps of:

dividing an interval defined as between a lowest and highest one of a plurality of predetermined data rates into m discriminating intervals, wherein m is an integer; and

calculating a difference between an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer and is less than m; and

when the difference between the average energies is greater than or equal to a threshold value, determining that the received signal in the (i+1)'th discriminating interval is transmitted at a data rate corresponding to the i'th discriminating interval.

2. The method as claimed in claim 1, wherein the threshold value is defined as $A^2/2$, wherein A represents a transmission power level of the received signal up to the i'th discriminating interval.

3. A device for detecting a data rate in a mobile communication system, in which an interval is defined as between a lowest and highest one of a plurality of given data rates being divided into m discriminating intervals, wherein m is an integer, the device comprising:

an energy calculator for calculating an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer and is less than m;

an energy differentiator for calculating a difference between the average energy of received signals up to the i'th discriminating interval and the average energy of received signals for the (i+1)'th discriminating interval; and

a data rate decision block for determining a data rate corresponding to the i'th discriminating interval as a data rate for the received signal in the (i+1)'th discriminating interval, when the difference between the average energies calculated in the energy calculator is greater than a threshold value.

4. The device as claimed in claim 3, wherein the threshold value is defined as $A^2/2$, wherein A represents a transmission power level of the received signal up to the i'th discriminating interval.

5. A method for detecting a data rate in a mobile communication system, in which a base station has previously provided a mobile station with information about a plurality of data rates variably serviceable and the mobile station detects one of the plurality of data rates as a data rate for a received signal, the method comprising the steps of:

(a) dividing an interval defined as between a lowest and highest one of the plurality of data rates into m discriminating intervals, wherein m is an integer; and

(b) calculating an average energy of a received signal corresponding to a first discriminating interval out of the m discriminating intervals;

(c) calculating an average energy of a received signal corresponding to a second discriminating interval next to the first discriminating interval;

(d) calculating a difference between the average energies obtained in steps (b) and (c); and

(e) estimating that the received signal for the second discriminating interval is transmitted at a data rate corresponding to the received signal for the first discriminating interval, when the difference between the average energies is greater than or equal to a threshold value, or

setting the first discriminating interval as the next discriminating interval when the difference between the average energies is greater than or equal to the threshold value,

the steps (b) to (e) for the received signals up to the newly set discriminating interval being repeated until the difference exceeds the threshold value.

FIG. 1

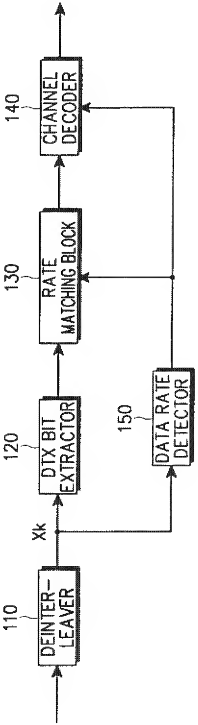
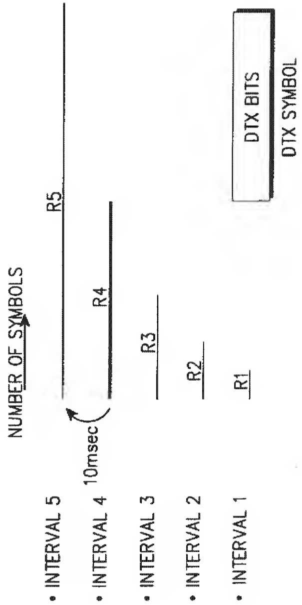

110
DEINTER-
LEAVER
Xk
120
DTX BIT
EXTRACTOR
130
RATE
MATCHING BLOCK
140
CHANNEL
DECODER
150
DATA RATE
DETECTOR

FIG. 2



• INTERVAL 5
• INTERVAL 4
• INTERVAL 3
• INTERVAL 2
• INTERVAL 1
10msec
NUMBER OF SYMBOLS
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R2
R3
R4
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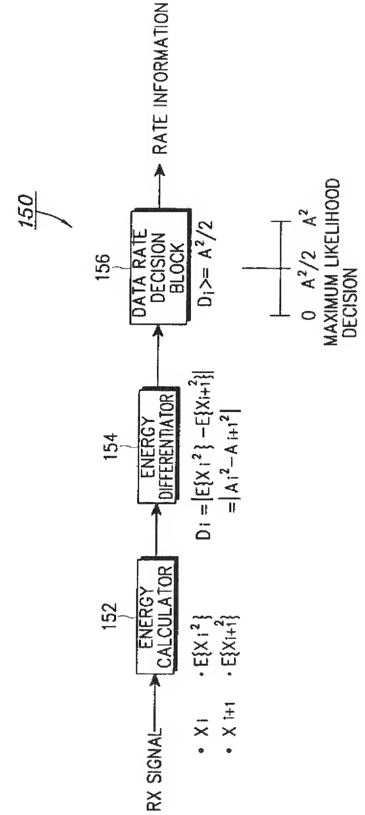
DTX SYMBOL

FIG. 3


150
RX SIGNAL
• X_i
• X_{i+1}
152
ENERGY CALCULATOR
• $E\{X_i^2\}$
• $E\{X_{i+1}^2\}$
154
ENERGY DIFFERENTIATOR
$D_i = |E\{X_i^2\} - E\{X_{i+1}^2\}|$
$= |A_i^2 - A_{i+1}^2|$
156
DATA RATE DECISION BLOCK
$D_i >= A^2/2$
RATE INFORMATION
0
$A^2/2$
A^2
MAXIMUM LIKELIHOOD DECISION

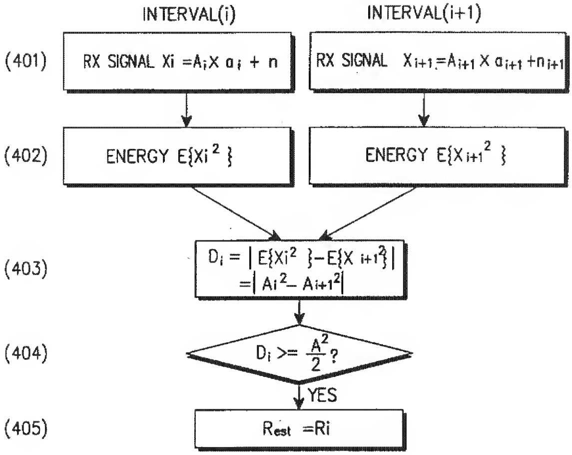

INTERVAL(i)
INTERVAL(i+1)
(401)
RX SIGNAL Xi =AiX ai + n
RX SIGNAL Xi+1=Ai+1 X ai+1 +ni+1
(402)
ENERGY E{Xi 2 }
ENERGY E{X i+1 2 }
(403)
Di = | E{Xi2 }−E{X i+12}|
=| Ai2− Ai+12|
(404)
Di >= A2/2 ?
YES
(405)
Rest =Ri

FIG. 4

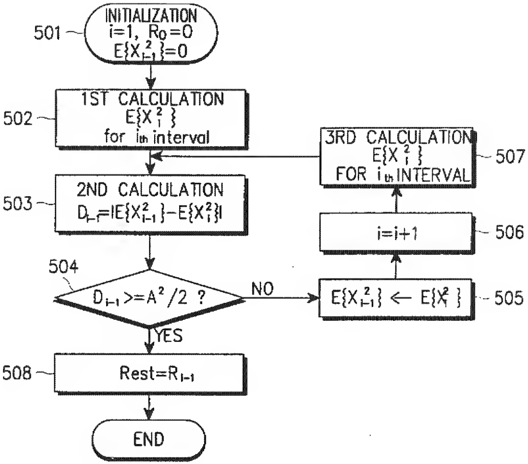

INITIALIZATION
i=1, R_0=0
$E\{X_{i-1}^2\}=0$
501
1ST CALCULATION
$E\{X_i^2\}$
for i_{th} interval
502
2ND CALCULATION
$D_{i-1}=|E\{X_{i-1}^2\}-E\{X_i^2\}|$
503
504
$D_{i-1}>=A^2/2$?
NO
YES
$E\{X_{i-1}^2\} \leftarrow E\{X_i^2\}$
505
i=i+1
506
3RD CALCULATION
$E\{X_i^2\}$
FOR i_{th} INTERVAL
507
Rest=R_{i-1}
508
END

FIG. 5

INTERNATIONAL SEARCH REPORT

International application No.
PCT/KR00/00740

A. CLASSIFICATION OF SUBJECT MATTER

IPC7 H04B 7/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimun documentation searched (classification system followed by classification symbols)

IPC7 H04B, H04L

Documentation searched other than minimun documentation to the extent that such documents are included in the fileds searched

Korean Patents and applications for inventions since 1975
Korean Utility models and applications for Utility models since 1975

Electronic data base consulted during the intertnational search (name of data base and, where practicable, search trerms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5566206 A (Qualcomm INC.) 15 October 1996 See the whole document	1,3,5
A	US 5671255 A (Motorola INC.) 23 September 1997 See the whole document	1,3,5
A	US 5751725 A (Qualcomm INC.) 12 May 1998 See the whole document	1,3,5

☐ Further documents are listed in the continuation of Box C.

☒ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevence
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevence; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevence; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents,such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
24 OCTOBER 2000 (24.10.2000)	25 OCTOBER 2000 (25.10.2000)

Name and mailing address of the ISA/KR	Authorized officer
Korean Industrial Property Office Government Complex-Taejon, Dunsan-dong, So-ku, Taejon Metropolitan City 302-701, Republic of Korea	YOON, Byoung Soo
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INTERNATIONAL SEARCH REPORT

Information on patent family members

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Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US 5566206 A	10. 15. 1996	WO 9501032 A1	05.01.1995
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		AU 4822097 A1	15.05.1998

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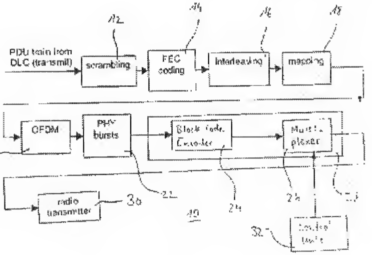
[Continued on next page]

(54) Title: MULTIPLEXING METHOD IN A MULTI CARRIER TRANSMIT DIVERSITY SYSTEM



(57) Abstract: Multiplexing Method in a Multi Carrier Transmit Diversity System The invention relates to a method of multiplexing data words in a multicarrier transmit diversity system. The method comprises the step of generating a plurality of data blocks, each data block comprising data words and each data word containing data symbols derived from a data signal, the step of determining for one or more data blocks in dependence on at least one transmission constraint if the data words of said one or more data blocks are to be multiplexed in the time domain or in the frequency domain and the step of multiplexing the data words of the data blocks in accordance with the determination result.

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before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

Multiplexing Method in a Multi Carrier Transmit Diversity System

BACKGROUND OF THE INVENTION

Technical Field

The present invention relates to the field of transmit antenna diversity and in particular to a method of multiplexing data words in a multi carrier transmit diversity system. The invention also relates to a multiplexer for multiplexing a sequence of data symbols and a demultiplexer for demultiplexing a multiplexed sequence of data symbols.

Discussion of the Prior Art

Peak transmission rates in wireless communication systems have steadily increased during the last years. However, peak transmission rates are still limited for example due to path loss, limited spectrum availability and fading.

Transmitter diversity is a highly effective technique for combating fading in wireless communications systems. Several different transmit diversity schemes have been proposed. In Li, Y.; Chuang, J.C.; Sollenberger, N.R.: Transmitter diversity for OFDM systems and its impact on high-rate data wireless networks, IEEE Journal on Selec. Areas. Vol. 17, No. 7, July 1999 the transmit diversity schemes of delay, permutation and space-time coding are examplarily described. According to the delay approach, a signal is transmitted from a first transmitter antenna and signals transmitted from further transmitter antennas are delayed versions of the signal from the first transmitter antenna. In the permutation scheme, the modulated signal is transmitted from a first transmitter antenna and permutations of the modulated signal

are transmitted from further transmitter antennas. By means of space-time coding a signal is encoded into several data words and each data word is transmitted from a different transmitter antenna. During transmission the data words are spread (or multiplexed) in the time domain by successively transmitting the data symbols of a data word over a single carrier frequency.

A further transmit diversity scheme for a multicarrier system is space-frequency coding. By means of space-frequency coding a signal is encoded into several data words and each data word is spread (or multiplexed) in the frequency domain by transmitting the data symbols of each data word on orthogonal frequencies, i.e. orthogonal subcarriers. An exemplary scheme for space-frequency coding is described in Mudulodu, S.; Paulraj, A.: A transmit diversity scheme for frequency selective fading channels, Proc. Globecom, San Francisco, pp. 1089-1093, Nov. 2000. According to the multicarrier system described in this paper, the data words relating to an encoded signal are preferably multiplexed in the time domain although orthogonal frequencies are available and the data words could thus also be multiplexed in the frequency domain. This is due to the fact that if multiplexing in the frequency domain is utilized the employed frequencies, i.e. subcarriers, must see the same channel, which may not always be possible in a frequency selective fading channel. However, in case the subcarriers experience the same channel, it is stated that either multiplexing in the time domain or multiplexing in the frequency domain or a combination of the two may be used. By combining multiplexing in the time domain and in the frequency domain the data symbols of a data word are simultaneously multiplexed in the time domain and in the frequency domain. This means that the data word is spread both across time and across frequencies.

Departing from the various transmit diversity schemes hitherto known there is a need for a method of multiplexing data

words in a multicarrier transmit diversity system which can easily be adapted to the specifications of different wireless communications systems. There is also a need for a corresponding multiplexer and a demultiplexer.

BRIEF DESCRIPTION OF THE INVENTION

The existing need is satisfied by a method of multiplexing data words in a multicarrier transmit diversity system which comprises the step of generating a plurality of data blocks, each data block comprising data words and each data word containing data symbols derived from a data signal, the step of determining for one or more of the data blocks in dependence on at least one transmission constraint if the data words of said one or more data blocks are to be multiplexed in a time domain or in a frequency domain and the step of multiplexing the data words of the data blocks in accordance with the result of the determination.

The multiplexing method of the invention is not restricted to a specific transmit diversity scheme as long as the utilized transit diversity scheme enables to generate from a data signal a plurality of data blocks having the above structure. For example, the transmit diversity schemes of block coding and of permutation allow to generate such data blocks. Preferably, the generated data blocks have the structure of a matrix similar to a space-time block code (STBC) matrix. Also, it is not required that the transmit diversity scheme guarantees full transmit diversity. In other words: the invention does not necessitate that each information symbol comprised within the data signal is transmitted from each transmitter antenna. Nonetheless, a preferred embodiment of the invention comprises the feature of full transmit diversity.

Moreover, the invention is not restricted to any number of transmit and receive antennas. Preferably, the number of data words per data block equals the number of transmit antennas

such that each data word of a data block may be transmitted from an individual transmitter antenna. If more than one receive antenna is provided, the receive diversity scheme of maximum-ratio combining can be applied. However, other receive diversity schemes may be used as well.

According to the invention, it is decided on a data block level how the data words are to be multiplexed. The decision on the data block level allows to change the multiplexing domain from one data block to a subsequent data block which is advantageous if one has to cope with specific predefined or varying transmission constraints. Also, the multiplexing method according to the invention can be applied in various wireless communication systems without major changes due to the specific multiplexing flexibility gained by selecting the multiplexing domain on the data block level. The multiplexing domain can be determined for each data block individually or simultaneously for a plurality of data blocks. For example, it can be decided for a sequence of data blocks that all data words comprised within the sequence of data blocks are to be multiplexed in either the time domain or in the frequency domain.

The multiplexing domain is determined by taking into account one or more transmission constraints. For example, the transmission constraints may comprise one or more physical transmission constraints or one or more data-related transmission constraints. It can also comprise both one or more physical transmission constraints and one or more data-related transmission constraints. The physical transmission constraints relate to the physical transmission conditions and can be derived from physical transmission parameters like a channel coherence bandwidth or a coherence time. The data-related transmission constraints relate to system specific constraints regarding for example the employed multiplexing scheme for the data words, the structure of the data signal,

the structure of the data blocks, the structure of the data words or the structure of the data symbols.

The data symbols may be derived from the data signal in various ways dependent on the transmit diversity scheme which is used. If, for example, the transmit diversity scheme of permutation is used, the data symbols contained in the data words are permutations of information symbols comprised within the data signal. As a further example, if the transmit diversity scheme of block coding is used, the data symbols contained in the data words are obtained from the information symbols comprised within the data signal by means of permutation and basic arithmetic operations, such as negation and complex conjugation.

The data signal from which the one or more data blocks are generated can have any format. According to a preferred embodiment, the data signal has the format of a sequence of discrete information symbols. For example, the data signal may have the structure of vectors, each vector comprising a predefined number of information symbols. The nature of the information symbols may depend on the specific wireless communication system in which the multiplexing method according to the invention is used. Many wireless communication systems employ different types of information symbols for different purposes. For example, some wireless communication systems use data signals which comprise a preamble, one or more user data sections or both a preamble and one or more user data sections. Usually, the preamble has a predefined structure and is utilized for purposes like channel estimation, frequency synchronization and timing synchronization.

In the following, several exemplary data-related transmission constraints are described in more detail. According to a first embodiment, the data-related transmission constraint is a predefined number N of data symbols to be comprised within each data word which is to be multiplexed in the time domain.

Usually, the number N of data symbols to be comprised within each data word cannot arbitrarily be chosen because it may depend for example on a code rate, on the condition that the data blocks have to be orthogonal matrices or on the availability of memory resources within the multicarrier transmit diversity system.

When the data words of a specific data block are to be multiplexed in the time domain, the number N of data symbols to be comprised within each data word may represent the number of time slots required for the transmission of a single data word over a single subcarrier. On the other hand, when the data words of a specific data block are to be multiplexed in the frequency domain, the number N of data symbols to be comprised within each data word stands for the number of subcarriers required to transmit a single data word during a single time slot.

Preferably, all data words of an individual data block comprise the same number of data symbols. If the data signal has such a structure that the number of data symbols comprised within each data word of a specific data block equals the predefined number N of data symbols, the data words of this data block may be multiplexed in the time domain. Otherwise, i.e. if the data signal has such a structure that the number of data symbols comprised within each data word of a specific data block does not equal the predefined number N of data symbols, the data words of this data block may be multiplexed in the frequency domain. Such a distinction will become necessary if the data signal or a portion thereof has a predefined length because the predefined length may imply that the total number N_S of data symbols which corresponds to the predefined length of the data signal or a portion thereof is not an integer multiple of the predefined number N of data symbols which should be comprised within a data word to be multiplexed in the time domain. In such a situation integer multiples of the predefined number N of data symbols are ar-

ranged in data blocks of data words which are multiplexed in a time domain and a remainder $N_R = \mathrm{mod}(N_D/N)$ of data symbols is arranged in a data block with data words which are multiplexed in the frequency domain.

Thus, by combining multiplexing in the time domain and in the frequency domain, data symbol fitting problems resulting from the predefined number N of data symbols to be comprised within each data word which is to be multiplexed in the time domain can be solved. Such data symbol fitting problems may for example become relevant when the data signal or a portion of the data signal has a predefined length because the wireless transmission system necessitates that the preamble portion or the user data portion of a data signal comprises a certain number of information symbols. Thus the data words of all data blocks except for the last data block are multiplexed in the time domain and the data words of the last data block are either multiplexed in the time domain or in the frequency domain depending on whether or not the data words of the last data block contain a number of data symbols which equals the predefined number N of data symbols.

So far the data-related transmission constraint of a predefined number N of data symbols to be comprised within each data word has been illustrated. According to a second embodiment, the data signal may comprise one or more periodic structures and the data related transmission constraint may be a preservation of the periodic structures such that the periodic structures are still periodic on a receiver side. The one or more periodic structures may be comprised within a preamble of the data signal, for example in the form of two or more identical preamble information symbols. Periodic structures are advantageous because they allow the use of synchronization algorithms with comparatively low complexity.

In case of multiplexing data symbols relating to periodic structures in the time domain the periodicity of the periodic

structures may get lost. Therefore, at least the data words of data blocks which relate to the periodic structures or parts of periodic structures are multiplexed in the frequency domain. By multiplexing the data words of these data blocks in the frequency domain it can be ensured that the periodicity of the periodic structures is maintained.

When the data words of data blocks generated from periodic structures or portions thereof are multiplexed in the frequency domain, the data words of data blocks generated from the remaining data signal are preferably multiplexed in the time domain. If, for example, the data words of data blocks generated from a preamble comprising periodic structures are multiplexed in the frequency domain, the data words of data blocks generated from a corresponding user data section may be multiplexed at least partly in the time domain.

Instead of data-related transmission constraints or in addition to data-related transmission constraints physical transmission constraints can be taken into account when deciding if the data words of one or more specific data blocks are to be multiplexed in the time domain or in the frequency domain. According to a preferred embodiment, the decision is made based on simultaneously evaluating a combination of one or more data-related transmission constraints and one or more physical transmission constraints.

The physical transmission constraints may be determined based on at least one of a channel coherence bandwidth

$$B_c \approx 1/\tau_{rms} \tag{1}$$

and a coherence time

$$t_c \approx 1/(2\ f_D) \tag{2}$$

wherein f_D is the doppler frequency and τ_{rms} is the root mean square of the delay spread of the channel impulse response.

Many transmit diversity schemes require constant or at least approximately constant channel parameters during transmission of one data word. If the data words are to be multiplexed in the frequency domain, a comparatively large coherence band width is required. This means that the relation

$$b_c \gg N/T \tag{3}$$

has to be fulfilled at least approximately, wherein N is the number of data symbols per data word and T is the duration of one of the data symbols, i.e. the duration of one time slot. A comparatively large coherence bandwidth requires that the channel parameters of N adjacent subcarriers have to be almost constant.

On the other hand, if the data words are to be multiplexed in the time domain, a comparatively large coherence time is required. This means that the relation

$$t_c \gg T \cdot N \tag{4}$$

has to be fulfilled at least approximately. In other words: N subsequent data symbols have to have nearly constant channel parameters, i.e. the channel parameters for a single subcarrier have to remain constant for a period of $N \cdot T$.

The physical transmission constraint may be determined by assessing if one or both of the relations (3) and (4) are fulfilled. Dependent on which of the two relations (3) and (4) is fulfilled best it is decided that the data words of the data blocks are to be multiplexed either in the time domain or in the frequency domain as a general rule. Deviations from this general rule may become necessary due to data-related

transmission constraints. For example, the data symbol fitting problem or the problem encountered with periodic structures may necessitate that although multiplexing in the time domain is generally to be preferred, the data words of at least some data blocks have to be multiplexed in the frequency domain. As a further example, changing transmission conditions may necessitate that the data words of some data blocks have to be multiplexed in the time domain and the data words of other data blocks have to be multiplexed in the frequency domain. As a third example, the data words of data blocks generated from a preamble may be multiplexed in the time domain and the data words of data blocks generated from a user data section may be multiplexed in the frequency domain. Such a combination has the advantage that the above-mentioned data symbol fitting problem, which usually is most relevant for the user data section, can be avoided while the multiplexing in the time domain of the data words of data blocks generated from the preamble allows a good channel estimation.

It was mentioned above that in order to achieve full diversity each information symbol has to be transmitted from each transmitter antenna. A further requirement of full transmit diversity is that the antenna signals are orthogonal to each other. This means that the data symbols have to be modulated onto subcarriers which are orthogonal to each other. However, the invention can also be practiced in case the subcarriers are not orthogonal.

BRIEF DESCRIPTION OF THE DRAWINGS

Further advantages of the invention will become apparent by reference to the following description of a preferred embodiment of the invention in the light of the accompanying drawings, in which :

Fig. 1 shows a data signal in the form of a physical burst to be processed in accordance with the invention;

Fig. 2 is a block diagram of a transceiver for wireless communication adapted to multiplex data words in accordance with the invention;

Fig. 3 shows several modulation schemes defined in the HIPERLAN/2 standard;

Fig. 4 shows the block code encoder of the transceiver depicted in Fig. 2;

Fig. 5 shows the configuration of a transmit antenna diversity scheme;

Fig. 6 is a schematic diagram of multiplexing data words in the time domain in accordance with the invention; and

Fig. 7 is a schematic diagram of multiplexing data words in the frequency domain in accordance with the invention.

DESCRIPTION OF PREFERRED EMBODIMENTS

Although the present invention can be used in any multicarrier transmit diversity system which employs a transmit diversity scheme allowing to generate data blocks having a structure as described above, the following description of preferred embodiments is examplarily set forth with respect to a multicarrier system which employs orthogonal frequency division multiplexing (OFDM) and which utilizes block coding for generating data blocks from a data signal.

The exemplary multicarrier system described below is derived from the European wireless local area network (WLAN) standard

high performance radio local area network type 2 (HIPERLAN/2). HIPERLAN/2 systems are intended to be operated in the 5 GHz frequency band. A system overview of HIPERLAN/2 is given in ETSI TR 101 683, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; System Overview, V1.1.1 (2000-02) and the physical layer of HIPERLAN/2 is described in ETSI TS 101 475; Broadband Radio Access networks (BRAN); HIPERLAN Type 2: Physical (PHY) Layer, V1.1.1 (2000-04). The multicarrier scheme of OFDM, which is specified in the HIPERLAN/2 standard, is very robust in frequency selective environments.

Up to now, the HIPERLAN/2 system and many other wireless communications systems do not support transmit diversity in spite of the fact that transmit diversity would improve the transmission performance and reduce negative effects of fast fading like Rayleigh fading. However, applying standard transmit diversity schemes to multicarrier communications systems may lead to various problems which are hereinafter examplarily described with respect to the HIPERLAN/2 system.

In Fig. 1 a typical physical burst of HIPERLAN/2 is illustrated. The physical burst comprises a preamble consisting of preamble symbols and a user data section consisting of user data symbols. In HIPERLAN/2 five different physical bursts are specified and each kind of physical burst has a unique preamble. However, the last three preamble symbols constitute a periodic structure which is identical for all preamble types. This periodic structure consists of a short OFDM symbol C32 of 32 samples followed by two identical regular OFDM symbols C64 of 64 samples. The short OFDM symbol C32 is a cyclic prefix which is a repetition of the second half of one of the C64 OFDM symbols. The so-called C-preamble depicted in Fig. 1 is used in HIPERLAN/2 for channel estimation, frequency synchronization and timing synchronization. The periodic structure within the C-preamble is necessary in order to allow the use of synchronization algorithms with comparatively low complexity.

The user data section of the physical burst depicted in Fig. 1 comprises a variable number N_{SYM} of OFDM symbols required to transmit a specific protocol data unit (PDU) train. Each OFDM symbol of the user data section consists of a cyclic prefix and a useful data part. The cyclic prefix consists of a cyclic continuation of the useful data part and is inserted before it. Thus, the cyclic prefix is a copy of the last samples of the useful data part. The length of the useful data part is equal to 64 samples and has a duration of 3,2 μs. The cyclic prefix has a length of either 16 (mandatory) or 8 (optional) samples and a duration of 0,8 μs or 0,4 μs, respectively. Altogether, a OFDM symbols thus has a length of either 80 or 72 samples corresponding to a symbol duration of 4,0 μs or 3,6 μs, respectively. An OFDM symbol therefore has an extension in the time domain. A OFDM symbol further has an extension in the frequency domain. According to HIPERLAN/2, a OFDM symbol extends over 52 subcarriers. 48 subcarriers are reserved for complex valued subcarrier modulation symbols and 4 subcarriers are reserved for pilots.

From the above it becomes clear that the HIPERLAN/2 physical burst depicted in Fig. 1 has a predefined length both in a time direction and in a frequency direction. Moreover, the physical burst of Fig. 1 comprises a periodic structure. It are among others these features of the physical burst of Fig. 1 which may lead to problems when the HIPERLAN/2 system or a similar wireless communication system has to be adapted to transmit diversity.

For typical HIPERLAN/2 scenarios the above relation (4) is usually fulfilled because the doppler frequency f_D is comparatively low. However, especially in outdoor environments, relatively large delay spreads can occur. Consequently, relation (3) cannot always be fulfilled. Therefore, a transmit diversity scheme like STBC multiplexing in the time domain should generally be a preferred transmit diversity scheme for

a HIPERLAN/2 scenario from the point of view that the channel over one space-time data word should be as constant as possible. However, severe problems arise when STBC is applied to physical bursts having the structure depicted in Fig. 1 or a similar structure.

Both the physical burst and the OFDM symbols comprised therein have predefined dimensions in the time domain and in the frequency domain. Concurrently, STBC requires that each STBC data word has a predetermined length N. Thus, data unit fitting problems arise if the dimension of e.g. an OFDM symbol of the preamble or of the user data section cannot be mapped on an integer multiple of the length of one STBC data word. Moreover, when applying STBC to the periodic C-preamble depicted in Fig. 1, the periodicity of the C-preamble gets lost. This is due to the fact that the one or more STBC data words relating to the second C64 OFDM symbol will no longer be equal to the one or more STBC data words relating to the first C64 OFDM symbol. The loss of periodicity, however, leads to the problem that the symbol synchronization algorithms which make use of a periodic structure within the preamble can no longer be employed. Also, the C32 OFDM symbol cannot serve any longer as a guard interval separating the OFDM symbols within the preamble. The reason therefore is that in case of multipath propagation the first C64 OFDM symbol interferes with the second C64 OFDM symbol which is no longer equal to the first C64 OFDM symbol.

The above problems and further problems not explicitly discussed above do not occur when the data words are multiplexed in accordance with the invention. In Fig. 2, the physical layer of a transceiver 10 which is adapted to implement the method according to the invention is illustrated. The transceiver 10 comprises a scrambler 12, an FEC coding unit 14, an interleaving unit 16, a mapping unit 18, an OFDM unit 20, a burst forming unit 22, a block code encoder 24, a multiplexer 26, a radio transmitter 30 and a control unit 32. The block

code encoder 24 and the multiplexer 25 together form an encoder/multiplexer unit 28.

The transceiver 10 depicted in Fig. 1 receives as input signal a PDU train from a data link control (DLC). Each PDU train consists of information bits which are to be framed into a physical burst, i.e. a sequence of OFDM symbols to be encoded, multiplexed and transmitted.

Upon receipt of a PDU train the transmission bit rate within the transceiver 10 is configured by choosing an appropriate physical mode based on a link adaption mechanism. A physical mode is characterized by a specific modulation scheme and a specific code rate. In the HIPERLAN/2 standard several different coherent modulation schemes like BPSK, QPSK, 16-QAM and optional 64-QAM are specified. Also, for forward error control, convolutional codes with code rates of 1/2, 9/16 and 3/4 are specified which are obtained by puncturing of a convolutional mother code of rate 1/2. The possible resulting physical modes are depicted in Fig. 3. The data rate ranging from 6 to 54 Mbit/s can be varied by using various signal alphabets for modulating the OFDM subcarriers and by applying different puncturing patterns to a mother convolutional code.

Once an appropriate physical mode has been chosen, the N_{BPDU} information bits contained within the PDU train are scrambled with the length-127 scrambler 12. The scrambled bits are then output to the FEC coding unit 14 which encodes the N_{BPDU} scrambled PDU bits according to the previously set forward error correction.

The encoded bits output by the FEC coding unit 14 are input into the interleaving unit 16 which interleaves the encoded bits by using the appropriate interleaving scheme for the selected physical mode. The interleaved bits are input into the mapping unit 18 where sub-carrier modulation by mapping the interleaved bits into modulation constellation points in ac-

cordance with the chosen physical mode is performed. As mentioned above, the OFDM subcarriers are modulated by using BPSK, QPSK, 16-QAM or 64-QAM modulation depending on the physical mode selected for data transmission.

The mapping unit 18 outputs a stream of complex valued subcarrier modulation symbols which are divided in the OFDM unit in groups of 48 complex numbers. In the OFDM unit a complex base band signal is produced by OFDM modulation as described in ETSI TS 101 475, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, V1.1.1 (2000-04).

The complex base band OFDM symbols generated within the OFDM unit 20, where pilot subcarriers are inserted, are input into the physical burst unit 22, where an appropriate preamble is appended to the PDU train and the physical burst is built. The physical burst produced by the physical burst unit 22 has a format as depicted in Fig. 1. The physical burst unit 22 thus outputs a sequence of complex base band OFDM symbols in the form of the physical burst to the block code encoder 24.

The function of the block code encoder 24 is now generally described with reference to Fig. 4. In general, the block code encoder 24 receives an input signal in the form of a sequence of vectors $X = [X_1X_2...X_K]^T$ of the length K. The block code encoder 24 encodes each vector X and outputs for each vector X a data block comprising a plurality of signal vectors $C^{(1)}, C^{(2)}, ..., C^{(M)}$ as depicted in Fig. 4. Each signal vector $C^{(1)}, C^{(2)}, ..., C^{(M)}$ corresponds to a single data word. Thus, the data block generated from the vector X comprises M data words wherein M is the number of transmitter antennas.

Each data word $C^{(i)}$ with $i = 1...M$ comprises N data symbols, i.e. each data word $C^{(i)}$ has a length of N. The value of N cannot be freely chosen since the matrix C spanned by the data words $C^{(i)}$ has to be orthogonal in this embodiment. Several examples for data blocks in the form of orthogonal code

matrices **C** are described in US 6,088,408. In the block coding approach described in the present embodiment all data symbols c_j^i of the code matrix **C** are derived from the components of the input vector **X** and are simple linear functions thereof or of its complex conjugate.

If a receive signal vector **Y** at one receive antenna is denoted by $\mathbf{Y} = [Y_1 Y_2 \ldots Y_N]^T$, the relationship between **Y** and the code matrix **C** is as follows:

$$\begin{bmatrix} Y_2 \\ \cdots \\ Y_N \end{bmatrix} = \begin{bmatrix} c_1 & \cdots & & c_1 \\ \cdots & & \cdots & \cdots \\ c_N^{(1)} & c_N^{(2)} & \cdots & c_N^{(M)} \end{bmatrix} \cdot \begin{bmatrix} h \\ \cdots \\ h^{(M)} \end{bmatrix} \qquad (5)$$

where $h^{(i)}$ represents the channel coefficient of the channel from the i-th transmit antenna to the receive antenna. A generalization to more receive antennas is straightforward.

In the following examples of possible block code matrices for two and three transmitter antennas, respectively, are discussed in more detail. The configuration of a wireless communication system with two transmit antennas and one receive antenna is depicted in Fig. 5. For two transmit antennas one possible block code matrix **C** with a code rate R = 1 is .

$$\mathbf{C} = \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_1^* \end{bmatrix} \qquad (6)$$

For three transmit antennas one possible block code matrix C with a code rate R = 0,5 is.

$$\mathbf{C} = \begin{bmatrix} -X_4 & -X_3 & X_2 \\ X_1^* & X_2^* & X_3^* \\ -X_2^* & X_1^* & -X_4^* \\ -X_3^* & X_4^* & X_1^* \\ -X_4^* & -X_3^* & X_2^* \end{bmatrix} \quad (7)$$

The code rate R is defined as the ratio of the length K of the input vector **X** and the length N of each code word $\mathbf{C}^{(i)}$:

$$R = K/N \quad (8)$$

As can be seen from Fig. 4, the block code encoder 24 outputs for each data signal in the form of a vector **X** a data block in the form of a matrix **C**. The data block output by the block code encoder 24 is input into the multiplexer 26 which multiplexes the data words (vectors $\mathbf{C}^{(i)}$) of each data block in accordance with an externally provided control signal either in the time domain or in the frequency domain. The control signal is generated by the control unit 32 based on an assessment of the transmission constraints. The assessment of the transmission constraints and the controlling of the multiplexer 26 by means of the control unit 32 will be described later in more detail.

In the multicarrier scheme OFDM, the output of the block code encoder 24 is modulated onto subcarriers which are orthogonal to each other. There exist essentially two possibilities to

multiplex a data block comprising individual data words in an OFDM system. According to a first possibility depicted in Fig. 6, the data words of a specific data block are extended in the time direction (STBC). In other words: The data words are multiplexed in the time domain. According to a second possibility, the data words of a data block are extended in the frequency direction as depicted in Fig. 7. This means that the data words are multiplexed in the frequency domain. Multiplexing the data words of a data block in the form of a code matrix in the frequency domain will in the following be referred to as space-frequency block coding (SFBC).

As can be seen from Figs. 6 and 7, the individual data words of a data block are transmitted from different transmit antennas. According to the multiplexing scheme of Fig. 6, an individual data block is transmitted on an individual subcarrier over a time interval of $N \cdot T$, wherein N is the number of data symbols per data word and T is the duration of one of the data symbols. According to the multiplexing scheme of Fig. 7, an individual data block is spread over N subcarriers and is transmitted during a time interval of T. It can clearly be seen that the multiplexing scheme of Fig. 6 can generally be employed when the relation (4) is fulfilled and the multiplexing scheme of Fig. 7 can generally be employed when the relation (3) is satisfied.

The encoded and multiplexed output signal of the encoder/multiplexer unit 28 is input into the radio transmitter 30. The radio transmitter 30 performs radio transmission over a plurality of transmit antennas by modulating a radio frequency carrier with the output signal of the encoder/multiplexer unit 28. The transceiver 10 of Fig. 2 further comprises a receiver stage not depicted in Fig. 2. The receiver stage has a physical layer with components for performing the inverse operations of the components depicted in Fig. 2. For example, the receiver stage comprises a descram-

bler, a FEC decoding unit, a demultiplexer/decoder unit with a demultiplexer and a block code decoder, etc.

Now, the control of the multiplexer 26 will be described in more detail with reference to both physical and data-related transmission constraints that may occur if physical bursts as the one depicted in Fig. 1 are employed. In accordance with typical HIPERLAN/2 scenarios, it is supposed that relation (4) is fulfilled and that it cannot always be guaranteed that relation (3) is fulfilled. This corresponds to the realistic situation that the basic performance of STBC transmission is better than the basic performance of SFBC transmission. Basic performance here means that only physical transmission constraints are taken into account. In such a case the control unit 32 may decide that the data blocks have to be multiplexed in the time domain. However, if the physical transmission parameters change, there might occur the case where relation (4) is no longer fulfilled whereas relation (3) is fulfilled at least approximately. In this case the control unit 32 will decide that the data words of the data blocks are no longer multiplexed in the time domain. Instead, the control unit 32 controls the multiplexer 26 such that the data words of the data blocks are multiplexed in the frequency domain.

So far only physical transmission constraints have been considered. Should data-related transmission constraints also be of importance, the control unit 32 controls the multiplexer 26 by additionally taking into account data-related transmission constraints.

It has been mentioned above that the transmission constraints which have to be considered in context with the physical burst depicted in Fig. 1 are the preservation of a periodic structure in the C-preamble and the provision of a predefined number N of data symbols in each data word which is to be

multiplexed in the time domain. These two data-related transmission constraints can occur in several combinations.

According to a first scenario, the data signal has the structure of the physical burst depicted in Fig. 1 and comprises a user data section and a preamble with a periodic structure. It is further supposed that the data-related transmission constraint of preserving the periodic structure has to be taken into account while no data symbol fitting problem occurs with respect to the user data section. In such a case the data words of data blocks relating to the preamble are multiplexed in accordance with SFBC in the frequency domain and the data words of data blocks relating to the user data section are multiplexed in accordance with STBC in the time domain. By multiplexing the data words derived from the preamble in the frequency domain a preservation of the order of the C32 OFDM symbols and the two C64 OFDM symbols can be achieved.

According to a second scenario derived from the physical burst depicted in Fig. 1, the periodic structure within the preamble has to be preserved and additionally the data symbol fitting problem has to be taken into account with respect to the user data section. Like in the first scenario, the data words of data blocks derived from the preamble are multiplexed in accordance with SFBC in the frequency domain. Due to the data symbol fitting problem the data words of the last data block relating to the user data structure contains less than the predefined number N of data symbols contained in data words of the previous data blocks. Therefore, only the data words (containing the predefined number N of data symbols) of the previous data blocks are multiplexed in accordance with STBC in the time domain. The data words of the last data block contain $N_R = mod(N_D/N)$ data symbols and are multiplexed in accordance with SFBC in the frequency domain, wherein N_D is the total number of data symbols to be transmitted over one transmit antenna.

According to a third scenario, the data-related transmission constraint of the preservation of a periodic structure within the preamble is not relevant but the data symbol fitting problem is relevant with respect to the user data section. In this case the data words of data blocks relating to the preamble are multiplexed in accordance with STBC in the time domain and the data words of data blocks relating to the user data section are multiplexed as described above for the second scenario. In other words: The data words of the last data block have a length of N_R data symbols and the data words of the previous data blocks have the predefined length of N data symbols.

According to a fourth scenario, the data-related transmission constraint of preserving a periodic structure has not to be taken into account and the physical transmission constraint of $B_c \gg N/T$ is at least approximately fulfilled. In this case the data words of data blocks relating to the preamble are multiplexed in accordance with STBC in the time domain and the data words of data blocks relating to the user data section are multiplexed in accordance with SFBC in the frequency domain. By using STBC for the preamble a good channel estimation can be performed. Due to the use of STBC for the preamble the slightly worse performance of SFBC can be compensated by means of receiver algorithms for interference suppression based on the good channel estimation. Using STBC for the preamble and SFBC for the user data section has the advantage that data symbol fitting problems with respect to the user data section do not appear.

Additional scenarios based on further combinations of data-related and physical transmission constraints can easily be realized in accordance with the invention. Also, the invention can easily be applied to data signals having a structure different from the structure of the physical burst depicted in Fig. 1. Although the invention is preferably practiced

with the transmit diversity scheme of a combination of STBC and SFBC, other transmit diversity schemes can be used as well.

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CLAIMS

1. A method of multiplexing data words in a multicarrier transmit diversity system, comprising:
   a) generating a plurality of data blocks (C), each data block (C) comprising data words ($C^{(i)}$) and each data word ($C^{(i)}$) containing data symbols ($c_j^{(i)}$) derived from a data signal;
   b) determining for one or more data blocks (C) in dependence on at least one transmission constraint if the data words ($C^{(i)}$) of said one or more data blocks (C) are to be multiplexed in the time domain or in the frequency domain; and
   c) multiplexing the data words ($C^{(i)}$) of the data blocks (C) in accordance with the determination in step b).

2. The method according to claim 1, wherein the data signal comprises at least one of a preamble and a user data section.

3. The method according to claim 1 or 2, wherein the at least one transmission constraint comprises a data-related transmission constraint.

4. The method according to claim 3, wherein the data-related transmission constraint is a pre-defined number (N) of data symbols ($c_j^{(i)}$) to be comprised within each data word ($C^{(i)}$) which is to be multiplexed in the time domain.

5. The method according to claim 4, wherein the data words ($C^{(i)}$) containing the predefined number (N) of data symbols ($c_j^{(i)}$) are multiplexed in the time domain and the data words ($C^{(i)}$) containing more or

less data symbols ($c_j^{(i)}$) are multiplexed in the frequency domain.

6. The method according to claim 4 or 5, wherein the data signal or a portion thereof has a predefined length and wherein integer multiples of the predefined number of data symbols ($c_j^{(i)}$) are arranged in data blocks (C) with data words ($C^{(i)}$) which are multiplexed in the time domain and a remainder of data symbols ($c_j^{(i)}$) is arranged in data blocks (C) with data words ($C^{(i)}$) which are multiplexed in the frequency domain.

7. The method according to claim 6, wherein the user data section of the data signal has the predefined length.

8. The method according to claim 7, wherein the data words ($C^{(i)}$) of data blocks (C) relating to the preamble are either multiplexed completely in the frequency domain or completely in the time domain dependent on the transmission constraint.

9. The method according to one of claims 1 to 8, wherein the data signal comprises one or more periodic structures (C32, C64).

10. The method according to claim 9, wherein the one or more periodic structures (C32, C64) are contained within the preamble.

11. The method according to claim 9 or 10, wherein the data-related transmission constraint is a preservation of the one or more periodic structures (C32, C64).

12. The method according to one of claims 9 to 11,

wherein at least the data words ($C^{(i)}$) of data blocks (C) relating to the periodic structures (C32, C64) are multiplexed in the frequency domain.

13. The method according to claim 12, wherein the data words ($C^{(i)}$) of data blocks (C) relating to the user data section are multiplexed in the time domain.

14. The method according to one of claims 1 to 13, wherein the at least one transmission constraint comprises a physical transmission constraint.

15. The method according to claim 14, wherein the physical transmission constraint is determined based on at least one of a coherence bandwidth and a coherence time.

16. The method according to claim 15, wherein the physical transmission constraint is determined by assessing if the relationship $B_c >> N/T$ is fulfilled, wherein B_c is the coherence bandwidth, N is the number of data symbols ($c_j^{(i)}$) per data word ($C^{(i)}$) and T is the duration of one of the data symbols ($c_j^{(i)}$).

17. The method according to claim 15 or 16, wherein the physical transmission constraint is determined by assessing if the relationship $t_c >> N \cdot T$ is fulfilled, wherein t_c is the coherence time, N is the number of data symbols ($c_j^{(i)}$) per data word ($C^{(i)}$) and T is the duration of one of the data symbols ($c_j^{(i)}$).

18. The method according to claim 16 or 17, wherein, when the physical transmission constraint $B_c >> N/T$ is at least approximately fulfilled, the data words ($C^{(i)}$) of data blocks (C) relating to the preamble are multiplexed in the time domain and the data words

($C^{(i)}$) of data blocks (C) relating to the user data sequence are multiplexed in the frequency domain.

19. The method according to one of claims 1 to 18, wherein the data blocks (C) are obtained from the data signal by means of block coding or by means of permutation.

20. The method according to one of claims 1 to 19, wherein the data symbols ($c_j^{(i)}$) are modulated onto subcarriers which are orthogonal to each other.

21. A multiplexer (26) adapted to multiplex data words in accordance with the method according to one of claims 1 to 20.

22. A demultiplexer adapted to demultiplex data words multiplexed by the multiplexer of claim 21.

23. A transceiver for wireless communication, comprising at least one of a multiplexer according to claim 21 and a demultiplexer according to claim 22.

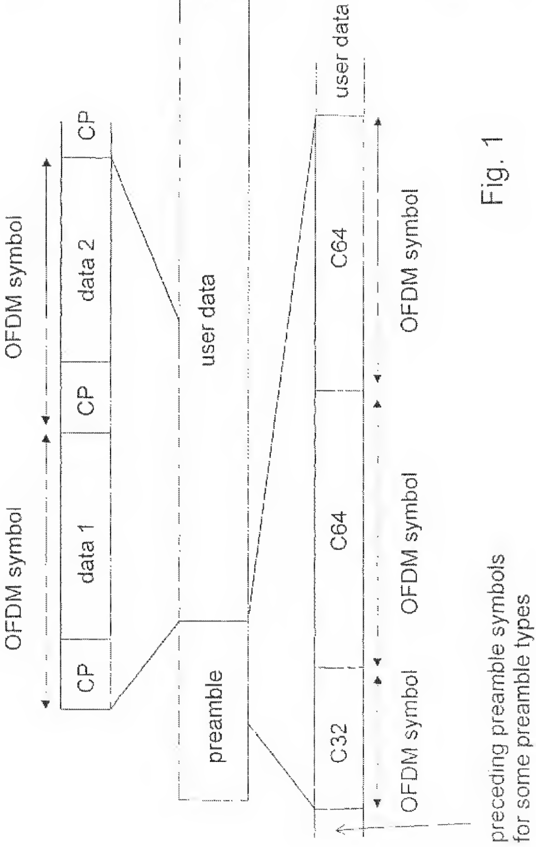
OFDM symbol
OFDM symbol
CP
data 1
CP
data 2
CP
preamble
user data
C32
C64
C64
user data
OFDM symbol
OFDM symbol
OFDM symbol
preceding preamble symbols
for some preamble types

Fig. 1

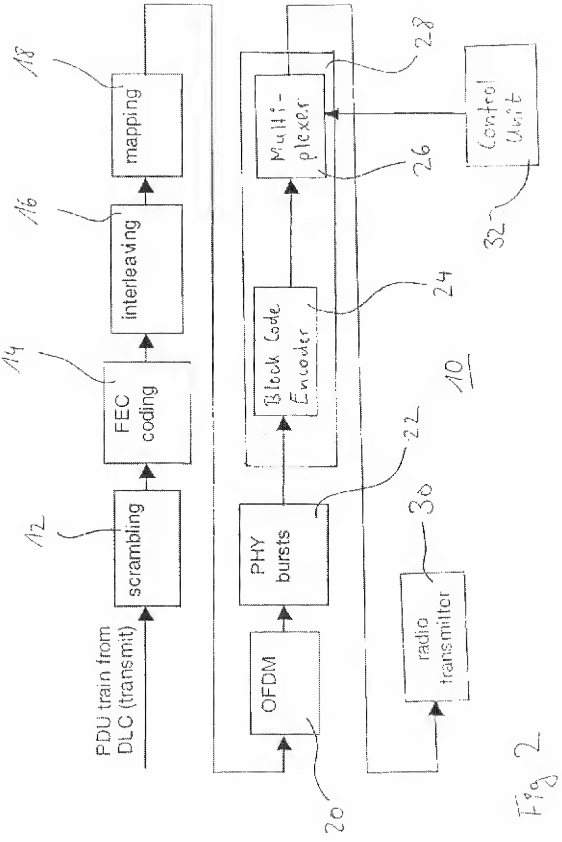
PDU train from DLC (transmit)
scrambling
FEC coding
interleaving
mapping
12
14
16
18
OFDM
PHY bursts
Block Code Encoder
Multi-plexer
20
22
24
26
28
Control Unit
32
radio transmitter
30
10

Fig 2

modulation scheme	code rate	bit rate
BPSK	1/2	6 Mbps
BPSK	3/4	9 Mbps
QPSK	1/2	12 Mbps
QPSK	3/4	18 Mbps
16-QAM	9/16	27 Mbps
16-QAM	3/4	36 Mbps
64-QAM	3/4	54 Mbps

Fig. 3

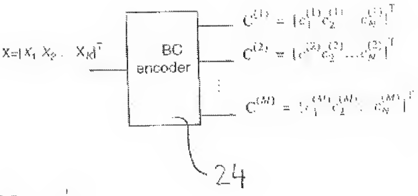

Fig. 4

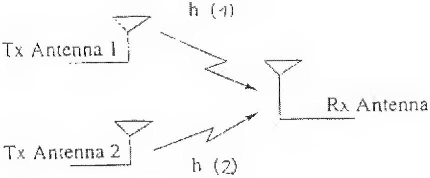

Fig. 5

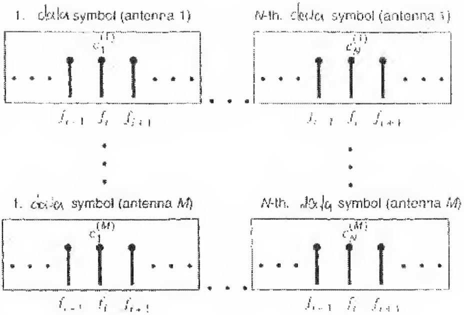
1. data symbol (antenna 1)
N-th. data symbol (antenna 1)
1. data symbol (antenna M)
N-th. data symbol (antenna M)

Fig. 6

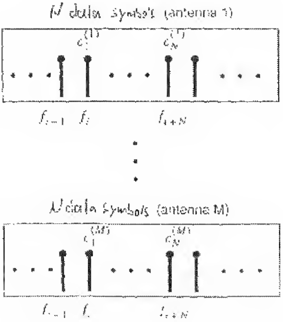
N data symbols (antenna 1)
N data symbols (antenna M)

Fig. 7

INTERNATIONAL SEARCH REPORT

International Application No
PCT/EP 02/02245

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L5/02 H04L1/00 H04B7/06 H04B7/12 H04L1/06
H04L1/04

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	MUDULODU S ET AL: "A transmit diversity scheme for frequency selective fading channels" GLOBECOM '00 - IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE. CONFERENCE RECORD (CAT. NO.00CH37137), PROCEEDINGS OF GLOBAL TELECOMMUNICATIONS CONFERENCE, SAN FRANCISCO, CA, USA, 27 NOV.-1 DEC. 2000, pages 1089-1093 vol.2, XP002172508 2000, IEEE, Piscataway, NJ, USA ISBN: 0-7803-6451-1 cited in the application the whole document	1-23

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[X] Further documents are listed in the continuation of box C.

[ ] Patent family members are listed in annex.

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"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
9 July 2002	17/07/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Toumpoulidis, T

INTERNATIONAL SEARCH REPORT

International Application No
PCT/EP 02/02245

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	LEE K F ET AL: "A space-frequency transmitter diversity technique for OFDM systems" GLOBECOM '00 - IEEE. GLOBAL TELECOMMUNICATIONS CONFERENCE. CONFERENCE RECORD (CAT. NO.00CH37137), PROCEEDINGS OF GLOBAL TELECOMMUNICATIONS CONFERENCE, SAN FRANCISCO, CA, USA, 27 NOV.-1 DEC. 2000, pages 1473-1477 vol.3, XP002172509 2000, Piscataway, NJ, USA, IEEE, USA ISBN: 0-7803-6451-1 the whole document -----	1-23
A	ALAMOUTI S M: "A simple transmit diversity technique for wireless communications" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. 16, no. 8, October 1998 (1998-10), pages 1451-1458, XP002100058 ISSN: 0733-8716 the whole document -----	1-23

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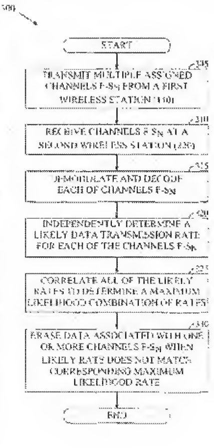
[Continued on next page]

(54) Title: METHOD AND APPARATUS FOR IS-95B REVERSE LINK SUPPLEMENTAL CODE CHANNEL FRAME VALIDATION AND FUNDAMENTAL CODE CHANNEL RATE DECISION IMPROVEMENT



(57) Abstract: The present invention provides a method and apparatus for maximizing throughput of a data call in a wireless communication system in which data is transmitted from a wireless station, such as a mobile station, on multiple assigned channels in accordance with a known transmission standard, such as IS-95B. The multiple assigned channels include a fundamental channel and at least one supplemental channel. Data is formatted into variable rate data frames and transmitted on the fundamental channel and the supplemental channel. A wireless receiver, such as a base station, receives the multiple assigned channels. The wireless receiver demodulates and decodes data frames associated with each of the multiple assigned channels. The wireless receiver determines a likely initial data rate for each demodulated and decode data frame. The wireless receiver correlates all of the likely data rates, by comparison to one another and to a relevant transmission protocol standard, to determine a maximum likelihood combination of data rates. The maximum likelihood combination of data rates includes a maximum likelihood data rate corresponding to each likely data rate. Decoded data frames are invalidated and erased when the likely data frame rates do not match corresponding maximum likelihood data rates.

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Published:

— *with international search report*

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METHOD AND APPARATUS FOR IS-95B REVERSE LINK SUPPLEMENTAL CODE CHANNEL FRAME VALIDATION AND FUNDAMENTAL CODE CHANNEL RATE DECISION IMPROVEMENT

BACKGROUND OF THE INVENTION

Field of the Invention

The present invention relates generally to wireless communication systems, and more particularly, to such a system for maximizing the useful data transmission throughput in a data call in which data is transmitted between wireless stations on multiple assigned channels.

Related Art

A wireless communication system can be used to transmit synchronous and asynchronous packet data between a wireless transmitter and a wireless receiver. For example, the wireless communication system can operate in accordance with a High Speed Packet Data (HSPD) feature of the "TIA/EIA/IS-95B Mobile Station-Base Station Compatibility Standard for Dual-mode Wideband Spread Spectrum Cellular Systems" (hereinafter referred to as IS-95B) to achieve a packet data transmission bandwidth of up to 115 kilobits-per-second (kbps). Under IS-95B, a mobile station can transmit data to a base station receiver on an IS-95B reverse-link traffic channel including a fundamental code channel (FCCH) and up to seven additional Supplemental Code Channels (SCCHs). The FCCH is a variable rate channel capable of operating at data transmission rates including a full rate, a half rate, a quarter rate, and an eighth rate. On the other hand, the SCCH operates only at a full rate when data is to be transmitted, and at a zero rate when no data is available.

Packet data transmitted on the FCCH and SCCHs is partitioned into 20 millisecond (ms) variable rate data frames. Although the data rate can change rapidly, for example, on a frame by frame basis, rate information is typically not included in each transmitted data frame for at least two reasons. First, including rate information in each data frame wastes data bandwidth, and second, corruption of such transmitted rate information would adversely affect the entire frame. Since rate information is not included in each transmitted data frame, the receiver must determine from each received data frame (without the aid of embedded rate information) the rate at which the frame was transmitted, to thereby enable the receiver to properly process the data in the data frame. Known methods of determining data frame rates exist for voice only traffic. However, such methods are insufficiently accurate and thus unsuitable for packet data traffic.

Therefore, there is a need in a variable rate communication system to accurately determine a transmitted data rate for packet data traffic at a receiver without embedding rate information into the transmitted data.

In the above described communication system, the mobile station sends signaling requests for SCCH assignment and de-assignment to the base station based on the amount of data the mobile station needs to transmit. In response, the base station dynamically allocates and de-allocates SCCHs via signaling messages. Assigning and de-assigning SCCHs via such signaling can be a relatively slow mechanism and thus wastes valuable data transmission bandwidth. For example, assigning or de-assigning an SCCH can take up to a half-second.

To reduce assignments and de-assignments and associated delays, a mobile station can operate in a discontinuous transmission (DTX) mode while a SCCH is assigned to the mobile station. The DTX mode permits the mobile station to stop transmitting on the assigned SCCH while data is unavailable. This is referred to as the DTX "black-out" period. The DTX mode also permits the mobile station to resume transmitting as soon as data becomes available, thus avoiding delays associated with assigning and de-assigning the SCCH.

Transmitted data frames typically do not include DTX "on/off" information for similar reasons as mentioned above with regard to rate information. Since the receiver of the assigned SCCH receives no explicit indicator regarding the black-out periods, the receiver continuously demodulates and decodes the SCCH as long as the SCCH is assigned, even during the black-out period when no data is being transmitted, that is, when the demodulated and decoded data is invalid.

Therefore, it is desirable at a receiver in a communication system to discriminate between data transmission periods and black-outs so as to reduce a likelihood that invalid data is declared to be valid at the receiver.

In accordance with IS-95B, each transmitted SCCH data frame includes a 12 bit Cyclic Redundancy Code (CRC) for checking the validity of the data in the data frame at the receiver. Additional observable metrics, such as a Yamamoto measure, a symbol error rate, a frame energy, and so on, can be used to further improve on the CRC check. There is a finite probability ($2^{-12} = 2.4 \times 10^{-4}$) that demodulated random data associated with the black-out period, or noise corrupting a received data frame, will cause an erroneous match of the 12 bit CRC. In the case of a black-out period, a non-existent SCCH data frame or "random frame" corresponding to the erroneous CRC match, erroneously labels the invalid random frame as a valid data frame.

As is known, the transmitter and receiver typically implement complementary or parallel, layered, communication protocol layers including a physical protocol layer and an overlaying Radio Link Protocol (RLP) layer. One known RLP layer useable in wireless data communication stations is the IS-707 Radio Link Protocol. The physical layer sends (and receives) supposedly valid data frames (for example, data frames passing the CRC check as mentioned above) to (and from) the RLP. The RLP at the receiver tracks RLP frame sequence numbers embedded in the data frames for purposes of errored frames re-transmission and control.

During black out-periods, it has been observed that passing random frames as valid data frames to the RLP causes the RLP to initiate error control

processes. This can occur on either the FCCH or SCCHs. For example, the RLP will reset and re-synchronize itself if the received sequence number, supposedly embedded in the random frame, is outside of a predetermined sequence number window (for example, 255) away from an expected sequence number. Alternatively, the RLP will request a retransmission of all of the data frames between the received and expected sequence numbers. In either case, the RLP error control processes disadvantageously reduce useful data throughput on the channel since most of the available bandwidth is utilized to re-sync the RLP or retransmit numerous data frames.

Therefore, there is a need to more accurately validate data frames at a receiver in a communication system, to thereby reduce the occurrence of such RLP error control processes and correspondingly increase channel bandwidth efficiency over conventional techniques.

SUMMARY OF THE INVENTION

The present invention provides a method and apparatus for maximizing throughput of a data call in a wireless communication system in which data is transmitted from a wireless station, such as a mobile station, on multiple assigned channels in accordance with a known transmission standard, such as IS-95B. In one embodiment, the multiple assigned channels include a fundamental channel and at least one supplemental channel. Data is formatted into variable rate data frames and transmitted on the fundamental and supplemental channels. A wireless receiver, such as a base station, receives the multiple assigned channels. The wireless receiver demodulates and decodes data frames associated with each of the multiple assigned channels. The wireless receiver determines a likely initial data rate for each demodulated and decode data frame. The wireless receiver correlates all of the likely data rates, by comparison to one another and to a relevant transmission protocol standard, to determine a maximum likelihood combination of data rates. The maximum likelihood combination of data rates

includes a maximum likelihood data rate corresponding to each likely data rate. Decoded data frames are invalidated and erased when the likely data frame rates do not match corresponding maximum likelihood data rates.

Features and Advantages

The present invention overcomes the above mentioned problems and represents an improvement over known rate determination and data validation techniques in a wireless data communication receiver.

The present invention accurately determines a variable transmitted data rate for packet data traffic at a wireless receiver without embedding rate information into the transmitted data.

The present invention advantageously reduces a likelihood that invalid data will be declared valid at the wireless receiver during both periods of data transmission and black-outs. More specifically, the present invention enhances the accuracies of rate determination and data validation at the receiver, and results in an increase in a traffic channel bandwidth efficiency over conventional techniques.

In a communication system including fundamental and supplemental channels operating in accordance with IS-95B, the present invention improves the accuracies of rate determination and data validation on the fundamental channel using supplemental channel signal quality measurements.

BRIEF DESCRIPTION OF THE FIGURES

The foregoing and other features and advantages of the invention will be apparent from the following, more particular description of the exemplary embodiments of the invention, as illustrated in the accompanying drawings.

FIG. 1 is a block diagram of an exemplary digital communications system 100 in which the present invention can be implemented.

FIG. 1A is an illustration of an exemplary transmit timing diagram of an FCCH and an exemplary transmit timing diagram of a concurrently assigned SCCH.

FIG. 2 is a block diagram of an exemplary transmit channel processor and a block diagram of an exemplary receive channel processor from FIG. 1.

FIG. 3 is an illustration of an exemplary high-level method of determining a maximum likelihood combination of rates used for validating decoded frames at a receiver of FIG. 1.

FIG. 4 is an illustration of a method corresponding to an exemplary embodiment of the present invention, wherein a receiver of FIG. 1 receives IS-95B reverse-link traffic channels.

FIG. 5 is an illustration of three exemplary timing diagrams (a), (b), and (c) corresponding respectively to an FCCH and two assigned SCCHs, and used to illustrate the method of FIG. 4.

FIG. 6 is a block diagram of an exemplary computer system on which the present invention can be implemented.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. 1 is a block diagram of an exemplary digital communications system 100 in which the present invention can be implemented. In an exemplary embodiment, system 100 is a CDMA cellular telephone system. However, it is to be understood that the present invention is applicable to other types of communication systems such as personal communications systems (PCS), wireless local loop, private branch exchange (PBX) or other known systems. The present invention is also applicable to systems using other well known transmission modulation schemes such as TDMA. System 100 includes a wireless transmitter 110 and a wireless receiver 120, each of which can be part of a base station (also known as a cell-site) or a mobile station. Communication from transmitter 110 to receiver 120 when receiver 120 is disposed in a mobile

station is known as the "forward link," and communication from transmitter 110 to receiver 120 when receiver 120 is disposed in a base station is known as the "reverse link." In the exemplary embodiment, transmitter 110 is disposed in a wireless station, such as the mobile station, and receiver 120 is disposed in the base station. Also, transmitter 110 and receiver 120 operate in accordance with IS-95B. The exemplary CDMA system operating in accordance with IS-95B allows for data communications between users over terrestrial links. The exemplary embodiment also applies to a CDMA system operating in accordance with *International Telecommunications Union wireless data communication* standards for third generation, International Mobile Telecommunications (IMT-2000).

Exemplary transmitter 110 includes a controller 130 for controlling the operation of transmitter 110 and for exchanging communication signaling information with receiver 120 to assign and de-assign communication channels during call-setup and tear-down, for example. Transmitter 110 includes a transmit channel processor 132 for performing transmit channel processing for one or more communication channels assigned to transmitter 110.

A data source 134 provides data 136 at variable data rates to transmitter 110. Data 136 can be synchronous or asynchronous packet data, as is known in the art. In turn, transmitter 110 formats data 136 into consecutive, variable rate data frames, each having an exemplary duration of 20 milliseconds. In the exemplary embodiment, an RLP processing component (not shown) at transmitter 110 and operating in accordance with TIA/EIA/IS-707 (referred to as "IS-707"), embeds consecutive frame sequence numbers in consecutive data frames for purposes of error correction and control. Then, transmit channel processor 132 further processes the data frames to prepare the data frames for wireless transmission to receiver 120, as will be further described below.

Transmitter 110 transmits the data frames to receiver 120 on a traffic channel 140 assigned to transmitter 110. In the exemplary embodiment, traffic channel 140 is a reverse link IS-95B traffic channel operating in a accordance

with the HSPD Feature of IS-95B. The IS-95B reverse link traffic channel 140 includes a fundamental code channel (FCCH) F, and can include up to seven additional supplemental code channels (SCCHs) S_0, S_1, S_2, S_3, S_4, S_5, S_6. The FCCH is a variable rate channel capable of operating at data frame rates (also referred to herein as "rates") including an FCCH full rate, a half rate, a quarter rate, and an eighth rate. The FCCH can carry data 136 from data source 134 and signaling information. Each of the assigned SCCHs S_0-S_6 can operate at only an SCCH full rate when data is to be transmitted and at a zero rate during DTX periods when no data is available to be transmitted. Under IS-95B, SCCHs S_0-S_7 can only transmit (at the SCCH full rate) when the FCCH is concurrently transmitting at the FCCH full rate. The present invention takes advantage of this IS-95B traffic channel restriction to improve the accuracies of determining FCCH frame rates and validating received data frames, as will be further described below.

In accordance with IS-95B, the above mentioned rates fall into two categories, namely, a first rate set RS1, and a second rate set RS2. RS1 includes the following rates:

1) FCCH rates of 9600 bps (the RS1 FCCH full rate), 4800 bps, 2400 bps, or 1200 bps; and

2) SCCH rates of 9600 bps (the RS1 SCCH full rate) or zero bps.

On the other hand, RS2 includes the following rates:

1) FCCH rates of 14,000 bps, 7200 bps, 3600 bps, and 1800 bps; and

2) SCCH rates of 14,000 bps or zero bps. It is to be understood that the present invention is applicable to communication systems having a greater or lesser number of data frame rates.

Still with reference to FIG. 1, receiver 120 includes a controller 150 for controlling the receiver and for exchanging signaling information with transmitter 110 to assign and de-assign traffic channels. Receiver 120 also includes a receive channel processor 152 for receiving traffic channel 140 and for processing received data frames so as to recover packet data 154, corresponding to packet

data 136, at transmitter 110. Receiver 120 delivers packet data 154 to a data sink 160. In the exemplary embodiment, receiver 120 and transmitter 110 both implement complementary RLP layers in accordance with IS-707. Controller 150 can include one or more controllers, and can encompass one or more processing functions of receive channel processor 152.

The above mentioned channel transmission requirements of the IS-95B HSPD feature are illustratively depicted in FIG. 1A. An exemplary transmit timing diagram (a) of the FCCH F and an exemplary transmit timing diagram (b) of a concurrently assigned SCCH S_i, are depicted in FIG. 1A. Timing diagram (a) is a plot of the FCCH transmitted rate (Rate) versus time, and timing diagram (b) is a plot of the SCCH S_i transmitted rate (Rate) versus time.

Referring to timing diagram (a), transmitter 110 transmits on the FCCH at the full, quarter, half, eighth, and full rates during consecutive portions 172, 174, 176, 178 and 180 of the timing diagram. A time interval 182 represents the duration of a single transmitted data frame, such as 20ms.

Referring to timing diagram (b), transmitter 110 concurrently transmits on SCCH S_i at the SCCH full rate during portions 190 and 192 respectively coinciding with portions 172 and 180 of timing diagram (a), in accordance with IS-95B. Conversely, transmitter 110 transmits on SCCH S_i at a zero rate (that is, transmitter 110 does not transmit) during portion 194 of timing diagram (b) coinciding with portions 174-178 of timing diagram (a). Portion 194 of timing diagram (b) corresponds to a black-out or DTX period on SCCH S_i. Also, it is to be understood the FCCH frames can be transmitted at the FCCH full rate while the SCCH S_i is at the zero rate.

FIG. 2 is a block diagram of exemplary transmit channel processor 132 of transmitter 110 and a block diagram of exemplary receive channel processor 152 of receiver 120. In transmit channel processor 132, a variable rate data framer 206 receives variable rate data 136, frames the variable rate data into variable data rate frames (also referred to herein as "frames"), and provides the frames to a cyclic redundancy code and tail bit generator 208, as applicable under IS-95B (for

example, only 9600 and 4800 bps FCCH frames and 9600 SCCH frames receive CRCs under IS-95B RS1). CRC generator 208 generates a set of CRC bits, such as 12 CRC bits, to provide for error detection at receiver 120. In addition, generator 208 appends a sequence of tail bits to each frame. In the exemplary embodiment, generator 208 generates the CRC and tail bits in accordance with IS-95B. Generator 208 provides a data frame to an encoder 210 for encoding the data as symbols for error correction and detection at receiver 120. In the exemplary embodiment, encoder 210 is a convolutional encoder. Encoder 210 provides encoded symbols to an interleaver 212. Interleaver 212 reorders the encoded symbols in accordance with a predetermined interleaving format. In the exemplary embodiment, interleaver 212 is a block interleaver, which is known in the art.

Interleaver 212 provides a reordered data frame to a modulator 214 for modulating the data frame for transmission. In the exemplary embodiment, modulator 214 is a CDMA modulator. Modulator 214 provides a modulated data frame to a transmitter module 216. Transmitter module 216 up-converts and amplifies the up-converted signal for transmission via an antenna 218. Transmitter module 216 transmits data frames to receiver 120 on traffic channel 140.

Receiver 120 receives traffic channel 140 via an antenna 220. Antenna 220 provides the received traffic channel to a plurality of parallel receive channel processors 152_1-152_n. Each of receive channel processors 152_1-152_n is assigned by receiver controller 150 to perform receive channel processing on a corresponding one of the received traffic channels F and S_0-S_n (also referred to herein as "F-S_n"). For example, receive channel processor 152_1 can be assigned to the FCCH, while the next receive channel processor 152_2 can be assigned to SCCH S_0, and so on. In this manner, receive processing for any one of the received channels F-S_n can be performed independently of the receive processing for any of the other received traffic channels.

Receive channel processor 152_1 performs receive channel processing as is now described. Receive antenna 220 provides received traffic channel 140 to a receiver module 222. Receiver module 222 down converts and amplifies the received traffic channel and provides a down converted and amplified received traffic channel to a demodulator 224, which demodulates the received channel. In the exemplary embodiment, demodulator 224 is a CDMA demodulator. In another embodiment, each of receive channel processors 152_1-152_n can share a single demodulator. Demodulator 224 provides a demodulated signal, namely, demodulated data frames, to de-interleaver 228. De-interleaver 228 re-orders demodulated data frame symbols in accordance with a predetermined format, as is known in the art.

De-interleaver 228 provides a re-ordered data frame to a decoder 230 for decoding the data frame. In the case where receive channel processor 152_1 is assigned to the FCCH, decoder 230 is preferably a multi-rate Viterbi decoder capable of decoding FCCH full rate, half rate, quarter rate and eighth rate received data frames associated with the FCCH, as is known in the art. In the case where receive channel processor 152_1 is assigned to an SCCH, S_i, decoder 230 need only decode full rate data frames since SCCH S_i can operate at only the SCCH full rate or the zero rate. As mentioned above, although the transmitted date frame rate can change on a frame by frame basis, rate information is typically not included in each transmitted data frame. Therefore, receiver 120 determines the transmitted rate for each received data frame to accurately decode and validate the data frame.

The decoding and CRC checking processes for a received FCCH frame are now described. In the exemplary embodiment, decoder 230 decodes symbols in the received FCCH frame for each of the four possible transmitted rates (that is, the FCCH full, half, quarter, and eighth rates) so as to provide four separately decoded frames, each of which is provided to a CRC check detector 232. Using conventional techniques, CRC check detector 232 determines whether the CRC bits for each of the four decoded frames are correct. CRC check detector 232

performs a CRC check for the CRC bits in each of the four decoded frames to determine at which of the full, half, quarter, or eighth rates the currently received frame was transmitted. As a result, in one embodiment, CRC check detector 232 provides four check bits, C_1, C_2, C_4, C_8, where the subscripts "1", "2", "4", and "8" respectively corresponding to the full rate, half rate, quarter rate, and eighth rate, and where a binary value of "1" for a given CRC check bit can indicate that the CRC check bits passed, while a binary value of "0" can indicate that the CRC bits failed.

In addition, decoder 230 provides decoded frame data to a Symbol Error Rate (SER) detector 234. Specifically, SER detector 234 receives decoded frame bits and an estimate of the received symbol data from decoder 230. As is known, SER detector 234 re-encodes and re-decodes the decoded bits, and compares them to the estimate of the received symbol data from decoder 230. The SER is a count of the number of discrepancies between the re-encoded symbol data and the received symbol data. Therefore, SER detector 234 generates four SER values: SER_1, SER_2, SER_3, and SER_4.

Furthermore, decoder 230 provides information to a Yamamoto check detector 236 for providing a confidence metric based on the difference between the selected path through a trellis and the next closest path through the trellis. The Yamamoto quality metric is well known in the art, and is further described, for example, in U.S. Pat. Nos. 5,710,784 and 5,872,775. While the CRC check is dependent on the bits in each of the four decoded frames, the Yamamoto check is dependent on the decoding process of receiver 120. Yamamoto detector 136, similar to detectors 232 and 234, provides four Yamamoto values for each of the four possible rates: Y_1, Y_2, Y_4, and Y_8. Although detectors 232, 234, 236 are shown as separate elements, the detectors can be incorporated within the hardware and/or software processes of decoder 230.

Receive channel processor 152_1 collectively provides the CRC check bits, SER values, and Yamamoto values from respective detectors 232, 234 and 236 to controller or control processor 150 as a data frame quality metric signal 240_1.

Data frame quality metric signal 240_1 is indicative of the quality (and thus validity) of decoded data corresponding to the data frame. Using data frame quality metric signal 240_1, control processor 150 determines at which of the four rates the currently received FCCH data frame was transmitted. For example, in the exemplary embodiment, the control processor selects a rate corresponding to a passed CRC and a favorable SER value.

Receive channel processor 152_1 also provides a decoded frame signal 242_1 to the control processor. Decoded frame signal 242_1 includes each of the separately decoded frames corresponding to the four different frame rates. Decoded frame signal 242_1 can be provided to a decoded data memory buffer so as to be accessible to the control processor.

The decoding and CRC checking processes performed on a received SCCH frame are similar to those processes described above for a received FCCH frame, as is now described. In the case where a receive channel processor (such as receive channel processor 152_2) is assigned to SCCH S_1, the associated decoder 230 decodes each received data frame at only the SCCH full rate. In this case, the assigned receive channel processor provides a single decoded SCCH data frame to control processor 150. Also, the receive channel processor provides the associated data frame quality metrics (for instance, the CRC, SER and Yamamoto values) associated with the decoded SCCH frame to control processor 150. Thus, in the case where multiple receive channel processors 152_1-152_n respectively process multiple receive channels F, S_0-S_n, the receive channel processors respectively provide data frame quality metrics signals 240_1-240_n and decoded frame signals 242_1-242_n to the control processor.

High-Level Method

Receiver controller 150 uses the above described signal quality metrics signals 240_1-240_n to initially determine current FCCH and SCCH frame rates and to initially validate the associated, decoded FCCH and SCCH frames. The

present invention then refines and thus improves the accuracy of such initial determinations, as is further described below.

FIG. 3 is an illustration of an exemplary high-level method 300 of determining a maximum likelihood combination of rates used for validating decoded frames at receiver 120, according to the present invention. Method 300 advantageously improves the likelihood of providing only valid received frames to subsequent processing stages such as the RLP processing layer and/or data sync 160. In doing so, method 300 reduces RLP error processing and correspondingly increases useful traffic channel bandwidth over other known methods, such as, for example, methods using only the above mentioned initial determinations.

Method 300 begins at a step 305 when transmitter 110 transmits data frames on multiple assigned traffic channels $F\text{-}S_n$. At a next step 310, receiver 120 receives traffic channels $F\text{-}S_n$. At a next step 315, receiver 120 demodulates, de-interleaves and decodes each of the received channels $F\text{-}S_n$ as described in connection with FIG. 2.

At a next step 320, a rate for each of the received channels $F\text{-}S_n$ is initially determined independent of the other received channels. Each determined or detected rate can be considered a "likely" rate because it may be incorrect if, for example, errors have corrupted the corresponding transmitted frame. In the exemplary embodiment, the likely rate for each SCCH is determined to be the SCCH full rate when the CRC check bits pass and the SER values are favorable for the decoded SCCH frame. When the likely rate is equal to the SCCH full rate, the associated SCCH decoded frame is assumed valid. On the other hand, when the likely rate is determined to be the zero rate, the associated SCCH data frame is assumed invalid.

In the exemplary embodiment, the likely rate for the FCCH is determined based on CRC check bits C_1, C_2, C_4, C_8, and SER values SER_1, SER_2, SER_3, and SER_4, provided that CRC check bits are available. Specifically, the likely FCCH rate is determined to be the one of the four possible rates corresponding to the one of the four decoded frames having a passing CRC and a favorable SER value.

The decoded frame associated with the selected likely rate is initially assumed valid.

At a next step 325, all of the likely rates determined at step 320 are correlated to produce a Maximum Likelihood (ML) combination of rates for the received traffic channels. The ML combination of rates includes an ML rate corresponding to each likely rate. Each such ML rate can be a probabilistically more accurate estimate of the transmitted rate than is the corresponding likely rate. This is because each likely rate is determined independent of the other traffic channels, whereas the ML rate is determined by correlating all of the independent likely rates. Correlating the independent likely rates adds relevant cross-channel rate information, such as traffic channel interdependencies, to each of the ML rate determinations, to thereby produce a probabilistically better rate estimate.

The correlation includes a comparison of each likely rate to each of the other likely rates. In addition, the correlation can include a comparison of the likely rates to a relevant set of rules, such as the traffic channel transmission requirements for the particular standard (for example, IS-95B) under which the traffic channels were transmitted. Such a comparison adds further relevant information to the process of generating the ML rates. A correlation in accordance with the exemplary embodiment is further described below in connection with FIG. 4.

At a next step 330, one or more of the likely rates determined at step 320 are compared or matched against corresponding ML rates in the ML combination of rates to determine whether to invalidate any of the decoded frames (such as the decoded FCCH frame) initially assumed valid in previous step 320.

Then, all of the decoded frames confirmed as valid in step 330 are provided to the next level of processing, such as the RLP processing layer and/or data sync 160. On the other hand, data frames invalidated at step 325 (and previous step 320) are "erased," that is, such invalidated frames are not provided

to the next level of processing. For example, the FCCH frame and one or more SCCH frames may be invalidated at step 330, based on the results from step 325.

Exemplary Method Embodiment

FIG. 4 is an illustration of a method 400 corresponding to the exemplary embodiment of the present invention, wherein receiver 120 receives reverse-link traffic channels operating in accordance with IS-95B. The principles embodied in exemplary method 400 also apply to any wireless data communication system operating in accordance with IMT-2000. The method steps of FIG. 4 are first described below, and then, a rationale supporting the method steps is provided. Steps 305, 310, 315, and 320 described above in connection with FIG. 3 are collectively represented in a single initial step 405 of method 400.

Next, at a decision step 410, it is determined whether the likely FCCH rate is at the FCCH full rate. If the likely FCCH rate is at the FCCH full rate, the decoded FCCH frame is assumed valid for use at the next processing stage, and flow proceeds to a step 415.

At step 415, decoded frames associated with received SCCHs are validated based on the respective likely rates of the decoded frames, as follows. First, the likely rates for the SCCHs (that is, the likely rate of each SCCH frame transmitted concurrently with the FCCH frame) are determined as described above. Then, decoded SCCH frames associated with full rates and zero rates are respectively assumed valid and invalid. Invalid SCCH frames are erased.

On the other hand, if at step 410 it is determined that the FCCH rate is at other than full rate, then flow proceeds to a next decision step 420. At decision step 420 it is determined whether at least two of the SCCHs have likely rates equal to the SCCH full rate. If at least two of the SCCHs have likely rates equal to the SCCH full rate, then flow proceeds to a step 425 where the decoded FCCH data frame (determined to be at other than the FCCH full rate at step 410) is

assumed invalid and erased. The SCCHs are validated in accordance with their respective likely rates as described in connection with step 415.

On the other hand, if at step 420 it is determined that less than two of the SCCHs are at the SCCH full rate, then flow proceeds to a step 430 where all of the concurrently received SCCH decoded frames are erased.

Decisional Analysis

The decisional logic embodied in method 400 is supported by a combination of the IS-95B requirements described above and by a probability analysis now described. The probability analysis considers two relevant probabilities. A first relevant probability P_e arises when an FCCH frame is transmitted at the FCCH full rate. In this case there is a finite probability, P_e, that the likely FCCH rate initially determined at step 410 will be erroneous, that is, the likely rate may be determined to be a rate other than the full rate (such as a half, quarter or eighth rate). This finite probability P_e is referred to as the "rate determination error for a full rate frame". The probability of detecting a full rate frame as other than a full rate frame, that is, the "probability of a rate determination error for a full rate frame" can be determined from Table 1 below. Table 1 is an excerpt from the TIA/EIA-IS-98B "Recommended Minimum Performance Standards for Dual Mode Wideband Spread Spectrum Cellular Mobile Stations" (referred to herein as "IS-98B"). Table 1 tabulates for the FCCH the minimum probabilities of rate determination error for IS-95B Rate Set 1 (RS1) and Rate Set 2 (RS2) full rate frames. RS1 full rate frames are assumed for the present discussion.

FCCH Rate	Min. Probability of Rate Detection Error at 1% FER (obtained from IS-98B)	
	RS1 Full	*RS2 Full*
Half	1.67×10^{-5}	1.67×10^{-5}
Quarter	1.41×10^{-4}	2.38×10^{-4}
Eighth	1.73×10^{-4}	2.73×10^{-4}

Table 1: Minimum Probabilities of Rate Determination Errors

Table 1 includes a first column listing FCCH rates, a second column listing error probabilities for RS1 FCCH full rate frames, and a third column listing error probabilities for RS2 FCCH full rate frames. Table 1 includes three rows respectively corresponding to half, quarter and eighth rates. The first row indicates the probability of erroneously detecting a full rate frame as a half rate frame. Similarly, the second row indicates the probability of erroneously detecting a full rate frame as a quarter rate frame, and so on.

The total minimum probability of a rate determination error (P_e) for detecting an RS1 FCCH full rate frame as other than a full rate frame is the addition of the error probabilities, from Table 1, of detecting the frame rate as one of the other three frame rates. In other words, the probability of erroneously detecting an FCCH frame transmitted at the full rate as other than the full rate is given by:

$$P_d = 1.67 \times 10^{-5} + 1.41 \times 10^{-4} + 1.73 \times 10^{-4} = 3.31 \times 10^{-4}$$

A second probability, P_c, relates to erroneously detecting an invalid received SCCH frame as a valid frame, for example, during a DTX period. As mentioned above, a transmitted SCCH data frame includes a 12 bit CRC. When the CRC passes at receiver 120, the corresponding SCCH frame is assumed valid.

It is to be understood that SER can also used for supplemental rate decisions, but that it is ignored here to simplify this probability analysis. Invalid frames can be received, for example, during a DTX period, when transmitted frames are corrupted with noise, or when transmitted frames are substantially attenuated during transmission. In such circumstances, there is the finite probability P_c of detecting a valid CRC at receiver 120 even though invalid data is being received and demodulated. The random probability P_c of a 12 bit CRC matching any random bit sequence at receiver 120 is $2.4x10^{-4}$. Further, assuming SCCH channels are statistically independent from each other for the purpose of calculating such a random probability, then the random probability P_{cc} of two SCCHs both passing CRCs is given by:

$$P_{cc} = P_c \text{ x } Pc\text{, where } P_c = 2.4x10^{-4}$$

$$\text{therefore } P_{cc} = 2.4x10^{-4} \text{ x } 2.4x10^{-4} = 5.96x10^{-8}$$

A comparison between P_{cc} and P_c reveals that $P_{cc} << P_c$, by several orders of magnitude. Since SCCH data frames can only be transmitted (at the SCCH full rate) when FCCH data frames are transmitted at the FCCH full rate under IS-95B, the probabilistic comparison P_{cc} vs. P_c definitively suggests the following conclusion: when a FCCH frame is detected at a rate other than the full rate (for example, at the half, quarter or eighth rate) and at the same time or concurrently (that is, for the same 20 ms frame interval) at least two SCCH data frames associated with two SCCHs are detected at the full rate, it is much more likely than not that the FCCH non full rate determination is erroneous and that the FCCH data frame was actually transmitted at the full rate. In other words, the initial FCCH non full-rate determination is most likely wrong, and therefore should be overruled.

Under such circumstances, it is likely the FCCH data frame is corrupted (or a DTX period is in progress) and probabilities dictate that it is safer to invalidate and erase the FCCH frame than it is to provide such a corrupted frame

to the RLP. Method 400 thus improves FCCH rate detection during HSPD calls by filtering-out invalid FCCH data frames in accordance with the result of the above described correlation between all of the received traffic channel rates, and the further comparison of the rates against the IS-95B transmission requirements.

The above described probabilistic comparison P_{ce} vs. P_e definitively suggests the FCCH non full-rate determination should be overruled when at least two SCCHs are at the full-rate. On the other hand, when only one SCCH is determined to be at the full rate, a relevant probabilistic comparison P_c (2.4×10^{-4}) vs. P_e (3.31×10^{-4}) is much less definitive since P_c and P_e are substantially the same, that is, within an order of magnitude of one another. Relative to the earlier probability comparison, this comparison suggests it is just as likely the FCCH data frame was transmitted at the FCCH full rate as it was not transmitted at the FCCH full rate when only one SCCH channel is detected at the SCCH full rate. Under such conditions, probability does not justify overruling a determination that the FCCH is not full-rate based on a single SCCH channel being full rate.

Therefore, in the exemplary embodiment, when the FCCH rate is not full rate and only one SCCH is full rate, the SCCH frame is invalidated/erased while the FCCH data frame is assumed valid and provided to the next processing stage. This approach is taken because experience has shown erasure of a valid SCCH data frame is less harmful than providing an invalid SCCH frame to the RLP.

Method 400 is now illustrated with reference to FIG. 5. FIG. 5 is an illustration of exemplary timing diagrams (a), (b) and (c) corresponding respectively to the FCCH and two assigned SCCHs. In diagrams (a), (b) and (c), the timing waveforms in solid line represent transmitted frame rates. At receiver 120, the detected rates (that is, the determined likely rates) are in accordance with the transmitted rates, except during a first frame interval 505 and a second frame interval 510 (depicted in timing diagram (a)), where respective erroneous likely rates 505' (timing diagram (c)) and 510' (timing diagram (a)) are depicted in dotted line.

During interval 505, while the FCCH rate is at the half rate, $SCCH_2$ is erroneously determined to be at the SCCH full rate (that is, the SCCH likely rate is equal to the SCCH full rate). Such a condition is not allowed under IS-95B. In this situation, method 400 invalidates and erases a decoded $SCCH_2$ frame associated with interval 505 in favor of the FCCH half rate detected during the same time interval.

During interval 510, while the FCCH is erroneously determined to be at the FCCH half rate, at least two concurrent SCCH full rate frames are detected, namely, full rate frames for $SCCH_1$ and $SCCH_2$. Such a condition is not allowed under IS-95B. In this situation, method 400 invalidates and erases the decoded FCCH frame in favor of the two SCCH full rate frames.

Table 2 below provides an exemplary illustration of the operation of method 400. Table 2 tabulates SCCH and FCCH frame erasure decisions in accordance with method 400 when up to four SCCH are assigned and received at receiver 120. The legend or key for interpreting Table 2 is as follows:

F = Full Rate; and

!F = not full rate (that is, Quarter. Half or Eighth Rate);

Fund	S1	S2	S3	S4	S5	S6	S7	Action
!F	F							Erase S1
!F	F	F						Erase F
!F	F	!F						Erase S1
!F	!F	F						Erase S2
!F	!F	!F						Erase S1.S2
!F	F	F	F					Erase F
!F	F	F	!F					Erase F, S3
!F	F	!F	F					Erase F.S2
!F	F	!F	!F					Erase S1
!F	!F	F	F					Erase F.S1
!F	!F	F	!F					Erase S2
!F	!F	!F	F					Erase S3
!F	!F	!F	!F					Erase S1,S2,S3
!F	F	F	F	F				Erase F
!F	F	F	F	!F				Erase F, S4
!F	F	F	!F	F				Erase F, S3
!F	F	F	!F	!F				Erase F,S3.S4
!F	F	!F	F	F				Erase F,S2
!F	F	!F	F	!F				Erase F,S2, S4
!F	F	!F	!F	F				Erase F.S2,S3
!F	F	!F	!F	!F				Erase S1
!F	!F	F	F	F				Erase F.S1
!F	!F	F	F	!F				Erase F,S1,S4
!F	!F	F	!F	F				Erase F.S1.S3
!F	!F	F	!F	!F				Erase S2
!F	!F	!F	F	F				Erase F.S1,S2
!F	!F	!F	F	!F				Erase S3
!F	!F	!F	!F	F				Erase S4
!F	!F	!F	!F	!F				Erase S1-S4

Table 2 : Example of new algorithm

Receiver 120 can perform specific features of the present invention using receiver controllers, which in effect comprise a computer system. Although communication-specific hardware can be used to implement the present invention, the following description of a general purpose computer system is provided for completeness. The present invention is preferably implemented in software. Alternatively, the invention may be implemented using hardware or a combination of hardware and software. Consequently, the invention may be implemented in a computer system or other processing system.

An example of such a computer system 600 is shown in FIG. 6. In the present invention, for example, the above described methods or processes execute on computer system 600. The computer system 600 includes one or more processors, such as processor 604. The processor 604 is connected to a communication infrastructure 606 (for example, a bus or network). Various software implementations are described in terms of this exemplary computer system. After reading this description, it will become apparent to a person skilled in the relevant art how to implement the invention using other computer systems and/or computer architectures.

Computer system 600 also includes a main memory 608, preferably random access memory (RAM), and may also include a secondary memory 610. The secondary memory 610 may include, for example, a hard disk drive 612 and/or a removable storage drive 614, representing a floppy disk drive, a magnetic tape drive, an optical disk drive, etc. The removable storage drive 614 reads from and/or writes to a removable storage unit 618 in a well known manner. Removable storage unit 618, represents a floppy disk, magnetic tape, optical disk, etc. which is read by and written to by removable storage drive 614. As will be appreciated, the removable storage unit 618 includes a computer usable storage medium having stored therein computer software and/or data.

In alternative implementations, secondary memory 610 may include other similar means for allowing computer programs or other instructions to be loaded into computer system 600. Such means may include, for example, a removable

storage unit 622 and an interface 620. Examples of such means may include a program cartridge and cartridge interface (such as that found in video game devices), a removable memory chip (such as an EPROM, or PROM) and associated socket, and other removable storage units 622 and interfaces 620 which allow software and data to be transferred from the removable storage unit 622 to computer system 600.

Computer system 600 may also include a communications interface 624. Communications interface 624 allows software and data to be transferred between computer system 600 and external devices. Examples of communications interface 624 may include a modem, a network interface (such as an Ethernet card), a communications port, a PCMCIA slot and card, etc. Software and data transferred via communications interface 624 are in the form of signals 628 which may be electronic, electromagnetic, optical or other signals capable of being received by communications interface 624. These signals 628 are provided to communications interface 624 via a communications path 626. Communications path 626 carries signals 628 and may be implemented using wire or cable, fiber optics, a phone line, a cellular phone link, an RF link and other communications channels.

In this document, the terms "computer program medium" and "computer usable medium" are used to generally refer to media such as removable storage drive 614, a hard disk installed in hard disk drive 612, and signals 628. These computer program products are means for providing software to computer system 600.

Computer programs (also called computer control logic) are stored in main memory 608 and/or secondary memory 610. Computer programs may also be received via communications interface 624. Such computer programs, when executed, enable the computer system 600 to implement the present invention as discussed herein. In particular, the computer programs, when executed, enable the processor 604 to implement the process of the present invention. Accordingly, such computer programs represent controllers of the computer

system 600. By way of example, in a preferred embodiment of the invention, the processes performed by receiver controller 150 can be performed by computer control logic. Where the invention is implemented using software, the software may be stored in a computer program product and loaded into computer system 600 using removable storage drive 614, hard drive 612 or communications interface 624.

In another embodiment, features of the invention are implemented primarily in hardware using, for example, hardware components such as application specific integrated circuits (ASICs). Implementation of the hardware state machine so as to perform the functions described herein will be apparent to persons skilled in the relevant art(s).

While various embodiments of the present invention have been described above, it should be understood that they have been presented by way of example, and not limitation. It will be apparent to persons skilled in the relevant art that various changes in form and detail can be made therein without departing from the spirit and scope of the invention.

The present invention has been described above with the aid of functional building blocks illustrating the performance of specified functions and relationships thereof. The boundaries of these functional building blocks have been arbitrarily defined herein for the convenience of the description. Alternate boundaries can be defined so long as the specified functions and relationships thereof are appropriately performed. Any such alternate boundaries are thus within the scope and spirit of the claimed invention. One skilled in the art will recognize that these functional building blocks can be implemented by discrete components, application specific integrated circuits, processors executing appropriate software and the like or any combination thereof. Thus, the breadth and scope of the present invention should not be limited by any of the above-described exemplary embodiments, but should be defined only in accordance with the following claims and their equivalents.

What is Claimed is:

1. A method of maximizing throughput of a data call in a wireless communication system in which data is transmitted from a wireless station on multiple assigned channels, comprising the steps of:

   a. receiving the multiple assigned channels;

   b. demodulating and decoding each of the multiple assigned channels;

   c. determining a likely data rate of each of the multiple assigned channels; and

   d. correlating all of the likely data rates to determine one or more Maximum Likelihood (ML) data rates each corresponding to a likely data rate.

2. The method of claim 1, further comprising the step of

   e. invalidating data associated with one of the multiple assigned channels when the likely data rate and a corresponding ML data rate of the one of the multiple assigned channels do not match.

3. The method of claim 1, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first data rate on the fundamental channel, and

wherein data can be transmitted at a second data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is more likely than not that data is being transmitted at the first data rate on the fundamental channel when a plurality of supplemental channels have likely data rates equal to the second data rate,

the method further comprising invalidating and erasing demodulated and decoded data associated with the fundamental channel when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) the plurality of supplemental channels have likely data rates equal to the second data rate.

4. The method of claim 3, wherein the multiple assigned channels collectively form an IS-95B reverse-link traffic channel, and

wherein the first data rate corresponds to a fundamental channel full rate and the second rate corresponds to a supplemental channel full rate,

the method further comprising invalidating and erasing demodulated and decoded data associated with each of the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the fundamental channel full rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the supplemental channel full rate.

5. The method of claim 4, further comprising the step of providing non-invalidated data to a radio link protocol processing layer.

6. The method of claim 1, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first non-zero data rate on the fundamental channel, and

wherein data can be transmitted at a second non-zero data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is approximately equally likely that data is being transmitted and that data is not being transmitted at the first data rate on the fundamental channel when only one of a plurality of supplemental channels has a likely data rate equal to the second data rate,

the method further comprising invalidating and erasing demodulated and decoded data associated with each of the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the second data rate.

7. The method of claim 1, wherein the data transmitted on the multiple assigned channels is formatted into data frames, and wherein step (b) comprises the steps of:

demodulating the data frames to produce demodulated data frames; and

de-interleaving the demodulated data frames to produce de-interleaved data frames;

8. The method of claim 7, further comprising the steps of:

decoding the de-interleaved data frames to produce decoded data frames; and

generating a signal quality signal indicative of a signal quality for each of the decoded data frames.

9 The method of claim 8, likely data rate of each of the decoded data frames based on a corresponding signal quality metric signal.

10. The method of claim 8, wherein each of the data frames includes a Cyclic Redundancy Code (CRC), and wherein the generating step comprises at least one of:

generating a CRC for each of the decoded data frames; and

generating a Symbol Error Rate (SER) for each of the decoded data frames.

11. The method of claim 10, wherein step (c) comprises determining a likely data rate of each of the data frames on each of the multiple assigned channels based on at least one of a CRC and an SER for each of the decoded data frames.

12. Apparatus for maximizing throughput of a data call in a wireless communication system in which data is transmitted by a wireless station to a receiver on multiple assigned channels, comprising:

receiving means for receiving the multiple assigned channels;

demodulating means and decoding means for respectively demodulating and decoding each of the multiple assigned channels;

determining means for determining a likely data rate of each of the multiple assigned channels; and

correlating means for correlating all of the likely data rates to determine a maximum likelihood combination of data rates.

13. The apparatus of claim 12, wherein the maximum likelihood combination of data rates includes a maximum likelihood data rate corresponding to each said likely data rate, the apparatus further comprising

invalidating means for invalidating data associated with one of the multiple assigned channels when the likely data rate of the one multiple assigned channel as determined by the determining means fails to match a corresponding maximum likelihood data rate determined by the correlating means.

14. The apparatus of claim 12, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first data rate on the fundamental channel, and

wherein data can be transmitted at a second data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is more likely than not that data is being transmitted at the first data rate on the fundamental channel when a plurality of supplemental channels have likely data rates equal to the second data rate,

the apparatus further comprising means for invalidating and erasing demodulated and decoded data associated with the fundamental channel when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and at the same time,

b) the plurality of supplemental channels have likely data rates equal to the second data rate.

15. The apparatus of claim 14, wherein the multiple assigned channels collectively form an IS-95B reverse-link traffic channel, and

wherein the first data rate corresponds to a fundamental channel full rate and the second rate corresponds to a supplemental channel full rate,

the apparatus further comprising means for invalidating and erasing demodulated and decoded data associated with the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the fundamental channel full rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the supplemental channel full rate.

16. The apparatus of claim 15, further comprising a radio link protocol processing layer and means for providing non-invalidated data to the radio link protocol processing layer.

17. The apparatus of claim 13, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first non-zero data rate on the fundamental channel, and

wherein data can be transmitted at a second non-zero data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is approximately equally likely that data is being transmitted and that data is not being transmitted at the first data rate on the fundamental channel when only one of a plurality of supplemental channels has a likely data rate equal to the second data rate,

the apparatus further comprising means for invalidating and erasing demodulated and decoded data associated with the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the second data rate.

18. The apparatus of claim 13, wherein the data transmitted on the multiple assigned channels is formatted into data frames, and wherein:

the demodulating means includes means for demodulating the data frames to produce demodulated data frames; and

the de-interleaving means includes means for de-interleaving the demodulated data frames to produce de-interleaved data frames.

19. The apparatus of claim 18, wherein

the decoding means include means for decoding the de-interleaved data frames to produce decoded data frames; and

generating means for generating a signal quality signal indicative of a signal quality for each of the decoded data frames.

20. The apparatus of claim 19, wherein the determining means includes means for determining a likely data rate of each of the decoded data frames based on a corresponding signal quality metric signal:

21. The apparatus of claim 19, wherein each of the data frames includes a Cyclic Redundancy Code (CRC), and wherein the generating means comprises at least one of:

means for generating a CRC for each of the decoded data frames; and

means for generating a Symbol Error Rate (SER) for each of the decoded data frames.

22. The apparatus of claim 21, wherein the determining means determines a likely data rate of each of the data frames on each of

the multiple assigned channels based on at least one of a CRC and an SER for each of the decoded data frames.

23. A computer program product comprising computer usable media having computer readable program code means embodied in the media for causing application programs to execute on a computer processor in a wireless communication device to maximize throughput of a data call in a wireless communication system in which data is transmitted by a wireless station to the wireless communication device on multiple assigned channels, the wireless communication device including receiving means for receiving the multiple assigned channels, and demodulating and decoding means for demodulating and decoding each of the multiple assigned channels, the computer readable program code means comprising:

a first computer readable program code means for causing the processor to determine a likely data rate of each of the multiple assigned channels; and

a second computer readable program code means for causing the processor to correlate all of the likely data rates to determine a maximum likelihood combination of data rates.

24. The computer program product of claim 23, further comprising a third computer readable program code means for causing the processor to invalidate data associated with one of the multiple assigned channels when the likely data rate of the one multiple assigned channel fails to match a corresponding maximum likelihood data rate.

25. The computer program product of claim 23, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first data rate on the fundamental channel, and

wherein data can be transmitted at a second data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is more likely than not that data is being transmitted at the first data rate on the fundamental channel when a plurality of supplemental channels have likely data rates equal to the second data rate,

the computer program product further comprising a third computer readable program code means for causing the processor to invalidate and erase demodulated and decoded data associated with the fundamental channel when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) the plurality of supplemental channels have likely data rates equal to the second data rate.

26. The computer program product of claim 25, wherein the multiple assigned channels collectively form an IS-95B reverse-link traffic channel, and

wherein the first data rate corresponds to a fundamental channel full rate and the second rate corresponds to a supplemental channel full rate.

the computer program product further comprising a fourth computer readable program code means for causing the processor to invalidate and erase demodulated and decoded data associated with each of the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the fundamental channel full rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the supplemental channel full rate.

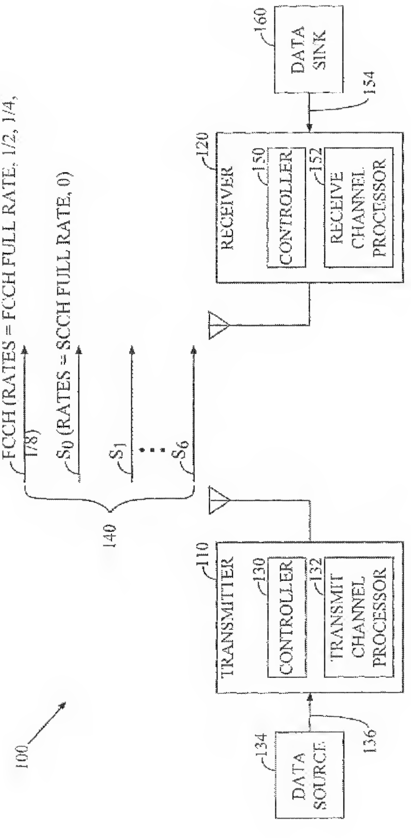
100
FCCH (RATES = FCCH FULL RATE, 1/2, 1/4, 1/8)
S0 (RATES = SCCH FULL RATE, 0)
S1
S6
140
110
TRANSMITTER
130
CONTROLLER
132
TRANSMIT CHANNEL PROCESSOR
134
DATA SOURCE
136
120
RECEIVER
150
CONTROLLER
152
RECEIVE CHANNEL PROCESSOR
154
160
DATA SINK

FIG. 1

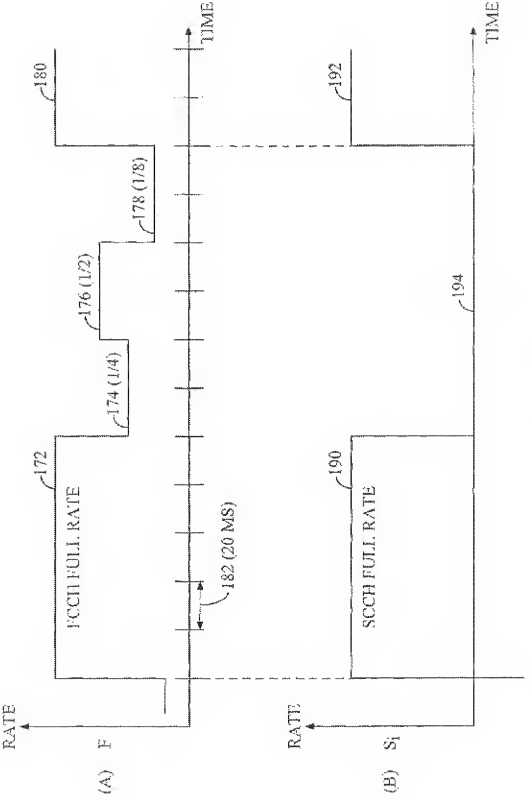
(A)
F
RATE
TIME
FCCH FULL RATE
172
174 (1/4)
176 (1/2)
178 (1/8)
180
182 (20 MS)
(B)
S_1
RATE
TIME
SCCH FULL RATE
190
192
194

FIG. 1A

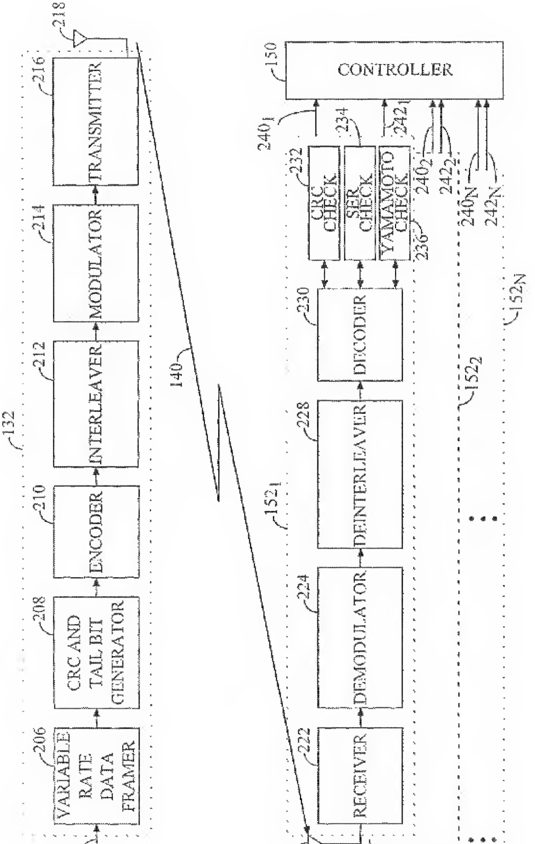
136
206
VARIABLE RATE DATA FRAMER
208
CRC AND TAIL BIT GENERATOR
210
ENCODER
212
INTERLEAVER
214
MODULATOR
216
TRANSMITTER
218
132
140
220
222
RECEIVER
224
DEMODULATOR
228
DEINTERLEAVER
230
DECODER
232
CRC CHECK
234
SER CHECK
236
YAMAMOTO CHECK
152_1
152_2
152_N
240_1
242_1
240_2
242_2
240_N
242_N
150
CONTROLLER

FIG. 2

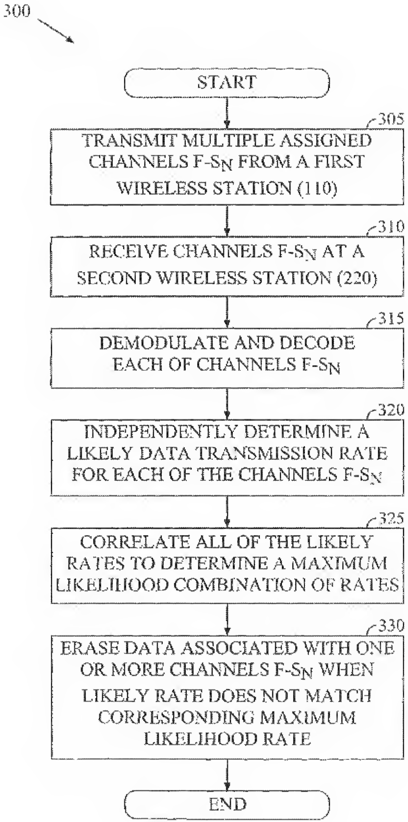
300
START
305
TRANSMIT MULTIPLE ASSIGNED CHANNELS F-S_N FROM A FIRST WIRELESS STATION (110)
310
RECEIVE CHANNELS F-S_N AT A SECOND WIRELESS STATION (220)
315
DEMODULATE AND DECODE EACH OF CHANNELS F-S_N
320
INDEPENDENTLY DETERMINE A LIKELY DATA TRANSMISSION RATE FOR EACH OF THE CHANNELS F-S_N
325
CORRELATE ALL OF THE LIKELY RATES TO DETERMINE A MAXIMUM LIKELIHOOD COMBINATION OF RATES
330
ERASE DATA ASSOCIATED WITH ONE OR MORE CHANNELS F-S_N WHEN LIKELY RATE DOES NOT MATCH CORRESPONDING MAXIMUM LIKELIHOOD RATE
END

FIG. 3

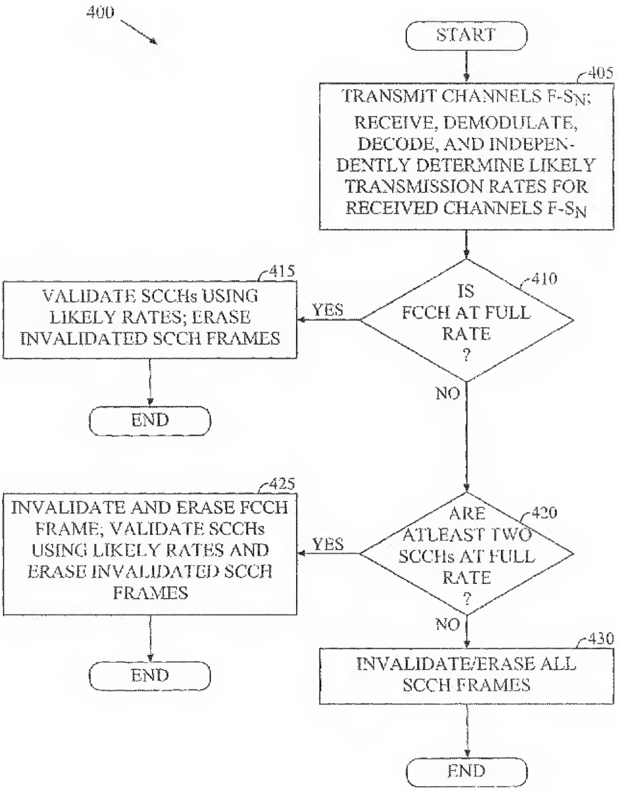
400
START
405
TRANSMIT CHANNELS F-S_N;
RECEIVE, DEMODULATE, DECODE, AND INDEPENDENTLY DETERMINE LIKELY TRANSMISSION RATES FOR RECEIVED CHANNELS F-S_N
410
IS FCCH AT FULL RATE ?
YES
415
VALIDATE SCCHs USING LIKELY RATES; ERASE INVALIDATED SCCH FRAMES
END
NO
420
ARE ATLEAST TWO SCCHs AT FULL RATE ?
YES
425
INVALIDATE AND ERASE FCCH FRAME; VALIDATE SCCHs USING LIKELY RATES AND ERASE INVALIDATED SCCH FRAMES
END
NO
430
INVALIDATE/ERASE ALL SCCH FRAMES
END

FIG. 4

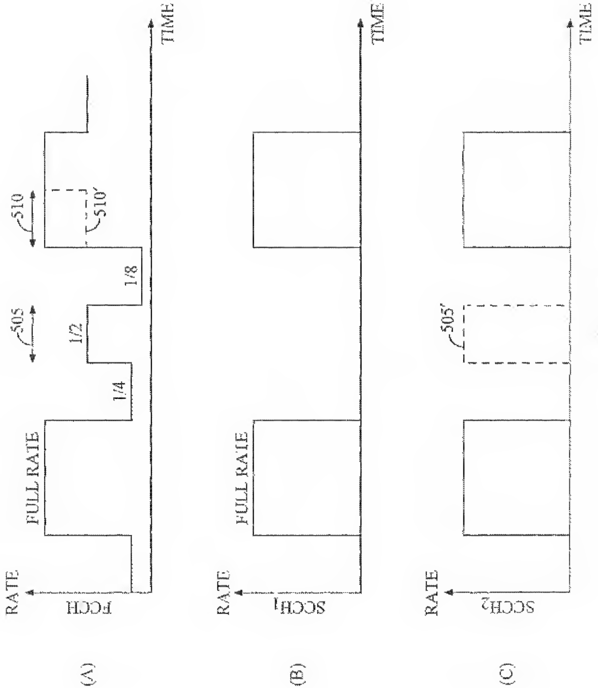
(A)
RATE
FCCH
FULL RATE
1/4
1/2
1/8
505
510
510′
TIME
(B)
RATE
$SCCH_1$
FULL RATE
TIME
(C)
RATE
$SCCH_2$
505′
TIME

FIG. 5

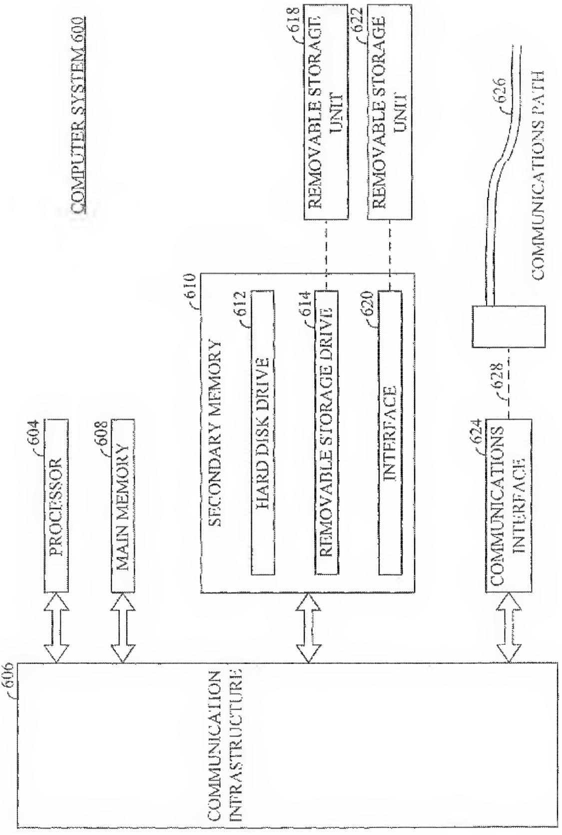
COMPUTER SYSTEM 600
604
PROCESSOR
608
MAIN MEMORY
610
SECONDARY MEMORY
612
HARD DISK DRIVE
614
REMOVABLE STORAGE DRIVE
620
INTERFACE
618
REMOVABLE STORAGE UNIT
622
REMOVABLE STORAGE UNIT
606
COMMUNICATION INFRASTRUCTURE
624
COMMUNICATIONS INTERFACE
628
626
COMMUNICATIONS PATH

FIG. 6

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/05417

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L25/02

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 6 108 372 A (LIN YU-CHUAN ET AL) 22 August 2000 (2000-08-22) figures 2,5	1,7-12, 23
Y	ANSI/TIA/EIA: "95-B Mobile Station - Base Station Compatibility Standard for Wideband Spread Spectrum Cellular Systems" 1 March 1999 (1999-03-01), TELECOMMUNICATIONS INDUSTRY ASSOCIATION, ARLINGTON XP002206171 page 6-46, line 14-27 footnote on page 6-47 page 7-75, line 6-8	1,7-12, 23

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

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"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
16 July 2002	01/08/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Horbach, C

Form PCT/ISA/210 (second sheet) (July 1992)

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	EP 0 920 160 A (MATSUSHITA ELECTRIC IND CO LTD) 2 June 1999 (1999-06-02) abstract column 4, line 50 -column 5, line 42; figures 4,5	1-26

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
US 6108372	A	22-08-2000	AU	5096498 A	22-05-1998
			EP	0935866 A1	18-08-1999
			JP	2001503233 T	06-03-2001
			WO	9819431 A1	07-05-1998
			US	2001055334 A1	27-12-2001
EP 0920160	A	02-06-1999	JP	11163961 A	18-06-1999
			EP	0920160 A2	02-06-1999
			US	6292920 B1	18-09-2001

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
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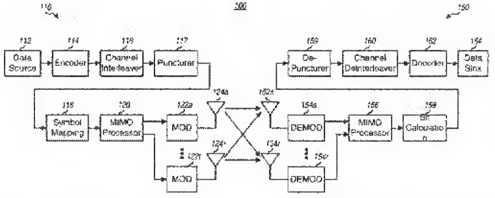

(57) Abstract: Coding techniques for a (e.g., OFDM) communication system capable of transmitting data on a number of transmission channels at different information bit rates based on the channels' achieved SNR. A base code is used in combination with common or variable puncturing to achieve different coding rates required by transmission channels. The data (i.e., information bits) for a data transmission is encoded with the base code, and the coded bits for each channel (or group of channels with the similar transmission capabilities) are punctured to achieve the required coding rate. The coded bits may be interleaved (e.g., to combat fading and remove correlation between coded bits in each modulation symbol) prior to puncturing. The unpunctured coded bits are grouped into non-binary symbols (e.g., using Gray mapping). The modulation symbol may be preconditioned and prior to transmission

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CODING SCHEME FOR A WIRELESS COMMUNICATION SYSTEM

BACKGROUND

I. Field

The present invention relates to data communication. More particularly, the present invention relates to a novel, flexible, and efficient coding scheme for encoding data for transmission on multiple transmission channels with different transmission capabilities.

II. Description of the Related Art

Wireless communication systems are widely deployed to provide various types of communication such as voice, data, and so on. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), orthogonal frequency division modulation (OFDM), or some other modulation techniques. OFDM systems may provide high performance for some channel environments.

In an OFDM system, the operating frequency band is effectively partitioned into a number of "frequency subchannels", or frequency bins. Each subchannel is associated with a respective subcarrier upon which data is modulated, and may be viewed as an independent "transmission channel". Typically, the data to be transmitted (i.e., the information bits) is encoded with a particular coding scheme to generate coded bits. For a high-order modulation scheme (e.g., QPSK, QAM, and so on), the coded bits are grouped into non-binary symbols that are then used to modulate the subcarriers.

The frequency subchannels of an OFDM system may experience different link conditions (e.g., different fading and multipath effects) and may achieve different signal-to-noise-plus-interference ratio (SNR). Consequently, the number of information bits per modulation symbol (i.e., the information bit rate) that may be transmitted on each subchannel for a particular level of performance may be different from subchannel to subchannel. Moreover, the link conditions typically vary with time. As a result, the supported bit rates for the subchannels also vary with time.

The different transmission capabilities of the frequency subchannels plus the time-variant nature of the capabilities make it challenging to provide an effective coding scheme capable of encoding the supported number of

information bits/modulation symbol to provide the required coded bits for the subchannels.

Accordingly, a high performance, efficient, and flexible coding scheme that may be used to encode data for transmission on multiple subchannels with different transmission capabilities is highly desirable.

SUMMARY

Various aspects of the present invention provides efficient and effective coding techniques for a communication system capable of transmitting data on a number of "transmission channels" at different information bit rates based on the channels' achieved SNR. A number of coding/puncturing schemes may be used to generate the required coded bits (i.e., the information, tail, and parity bits, if a Turbo code is used). In a first coding/puncturing scheme, a particular base code and common puncturing is used for all transmission channels (e.g., all frequency subchannels in an OFDM system, or spatial subchannels of all frequency subchannels in an OFDM system with multiple input/multiple output antennas (MIMO), as described below). In a second coding/puncturing scheme, the same base code but variable puncturing is used for the transmission channels. The variable puncturing can be used to provide different coding rates for the transmission channels. The coding rate for each transmission channel is dependent on the information bit rate and the modulation scheme selected for the channel.

An embodiment of the invention provides a method for preparing data for transmission on a number of transmission channels in a communication system, e.g., an orthogonal frequency division modulation (OFDM) system. Each transmission channel is operable to transmit a respective sequence of modulation symbols. In accordance with the method, the number of information bits per modulation symbol supported by each transmission channel is determined (e.g., based on the channel's SNR). A modulation scheme is then identified for each transmission channel such that the determined number of information bits per modulation symbol is supported. Based on the supported number of information bits per modulation symbol and the identified modulation scheme, the coding rate for each transmission channel is determined. At least two transmission channels are associated with different coding rates because of different transmission capabilities.

Thereafter, a number of information bits is encoded in accordance with a particular encoding scheme to provide a number of coded bits. If a Turbo code

is used, a number of tail and parity bits are generated for the information bits (the coded bits include the information bits, tail bits, and parity bits). The coded bits may be interleaved in accordance with a particular interleaving scheme. For ease of implementation, the interleaving may be performed prior to puncturing. The coded bits (e.g., the tail and parity bits, if a Turbo code is used) are then punctured in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the transmission channels. The puncturing is adjusted to achieve different coding rates needed by the transmission channels. As an alternative, the puncturing may also be performed prior to interleaving.

Non-binary symbols are then formed for the transmission channels. Each non-binary symbol includes a group of interleaved and unpunctured coded bits and is mapped a respective modulation symbol. The specific number of coded bits in each non-binary symbol is dependent on the channel's modulation scheme. For a multiple-input multiple-output (MIMO) system capable of transmitting on a number of spatial subchannels for each frequency subchannel, the modulation symbols for each frequency subchannel may be pre-conditioned prior to transmission, as described below.

The invention provides methods and system elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system capable of implementing various aspects and embodiments of the invention;

FIG. 2 is a diagram that graphically illustrates an OFDM transmission from one of N_T transmit antennas in the MIMO system;

FIGS. 3A and 3B are diagrams of a parallel concatenated convolutional encoder;

FIG. 3C is a diagram of an embodiment of a puncturer and multiplexer, which may be used to provide variable puncturing of coded bits;

FIGS. 4A and 4B are flow diagrams of two coding/puncturing schemes for generating the required coded bits for a data transmission, which utilize a particular base code but common and variable puncturing schemes, respectively;

FIG. 5 is a diagram of a signal constellation for 16-QAM and a specific Gray mapping scheme;

FIG. 6 is a block diagram of an embodiment of a MIMO processor;

FIG. 7 is a block diagram of an embodiment of a system capable of providing different processing for different transmissions; and

FIG. 8 is a block diagram of an embodiment of the decoding portion of a receiving system.

DETAILED DESCRIPTION OF THE SPECIFIC EMBODIMENTS

FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system 100 capable of implementing various aspects and embodiments of the invention. Communication system 100 can be designed to implement the coding schemes described herein. System 100 can further be operated to employ a combination of antenna, frequency, and temporal diversity to increase spectral efficiency, improve performance, and enhance flexibility. Increased spectral efficiency is characterized by the ability to transmit more bits per second per Hertz (bps/Hz) when and where possible to better utilize the available system bandwidth. Improved performance may be quantified, for example, by a lower bit-error-rate (BER) or frame-error-rate (FER) for a given link signal-to-noise-plus-interference ratio (SNR). And enhanced flexibility is characterized by the ability to accommodate multiple users having different and typically disparate requirements. These goals may be achieved, in part, by employing a high performance and efficient coding scheme, multi-carrier modulation, time division multiplexing (TDM), multiple transmit and/or receive antennas, other techniques, or a combination thereof. The features, aspects, and advantages of the invention are described in further detail below.

As shown in FIG. 1, communication system 100 includes a first system 110 in communication with a second system 150. Within system 110, a data source 112 provides data (i.e., information bits) to an encoder 114 that encodes the data in accordance with a particular coding scheme. The encoding increases the reliability of the data transmission. The coded bits are then provided to a channel interleaver 116 and interleaved (i.e., reordered) in

accordance with a particular interleaving scheme. The interleaving provides time and frequency diversity for the coded bits, permits the data to be transmitted based on an average SNR for the subchannels used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol, as described below. The interleaved bits are then punctured (i.e., deleted) to provide the required number of coded bits. The encoding, channel interleaving, and puncturing are described in further detail below. The unpunctured coded bits are then provided to a symbol mapping element 118.

In an OFDM system, the operating frequency band is effectively partitioned into a number of "frequency subchannels" (i.e., frequency bins). At each "time slot" (i.e., a particular time interval that may be dependent on the bandwidth of the frequency subchannel), a "modulation symbol" may be transmitted on each frequency subchannel. As described in further detail below, the OFDM system may be operated in a MIMO mode in which multiple (N_T) transmit antennas and multiple (N_R) receive antennas are used for a data transmission. The MIMO channel may be decomposed into N_C independent channels, with $N_C \leq N_T$ and $N_C \leq N_R$. Each of the N_C independent channels is also referred to as a "spatial subchannel" of the MIMO channel, which corresponds to a dimension. In the MIMO mode, increased dimensionality is achieved and N_C modulation symbols may be transmitted on N_C spatial subchannels of each frequency subchannel at each time slot. In an OFDM system not operated in the MIMO mode, there is only one spatial subchannel. Each frequency subchannel/spatial subchannel may also be referred to as a "transmission channel". The MIMO mode and spatial subchannel are described in further detail below.

The number of information bits that may be transmitted for each modulation symbol for a particular level of performance is dependent on the SNR of the transmission channel. For each transmission channel, symbol mapping element 118 groups a set of unpunctured coded bits to form a non-binary symbol for that transmission channel. The non-binary symbol is then mapped to a modulation symbol, which represents a point in a signal constellation corresponding to the modulation scheme selected for the transmission channel. The bit grouping and symbol mapping are performed for all transmission channels, and for each time slot used for data transmission. The modulation symbols for all transmission channels are then provided to a MIMO processor 120.

Depending on the particular "spatial" diversity being implemented (if any), MIMO processor 120 may demultiplex, pre-condition, and combine the received modulation symbols. The MIMO processing is described in further detail below. For each transmit antenna, MIMO processor 120 provides a stream of modulation symbol vectors, one vector for each time slot. Each modulation symbol vector includes the modulation symbols for all frequency subchannels for a given time slot. Each stream of modulation symbol vectors is received and modulated by a respective modulator (MOD) 122, and transmitted via an associated antenna 124.

In the embodiment shown in FIG. 1, receiving system 150 includes a number of receive antennas 152 that receive the transmitted signals and provide the received signals to respective demodulators (DEMOD) 154. Each demodulator 154 performs processing complementary to that performed at modulator 122. The demodulated symbols from all demodulators 154 are provided to a MIMO processor 156 and processed in a complementary manner as that performed at MIMO processor 120. The received symbols for the transmission channels are then provided to a bit calculation unit 158 that performs processing complementary to that performed by symbol mapping element 118 and provides values indicative of the received bits. Erasures (e.g., zero value indicatives) are then inserted by a de-puncturer 159 for coded bits punctured at system 110. The de-punctured values are then deinterleaved by a channel deinterleaver 160 and further decoded by a decoder 162 to generate decoded bits, which are then provided to a data sink 164. The channel deinterleaving, de-puncturing, and decoding are complementary to the channel interleaving, puncturing, and encoding performed at the transmitter.

FIG. 2 is a diagram that graphically illustrates an OFDM transmission from one of N_T transmit antennas in a MIMO system. In FIG. 2, the horizontal axis represents time and the vertical axis represents frequency. In this specific example, the transmission channel includes 16 frequency subchannels and is used to transmit a sequence of OFDM symbols, with each OFDM symbol covering all 16 frequency subchannels. A time division multiplexing (TDM) structure is also illustrated in which the data transmission is partitioned into time slots, with each time slot having a particular duration. For the example shown in FIG. 2, the time slot is equal to the length of one modulation symbol.

The available frequency subchannels may be used to transmit signaling, voice, packet data, and so on. In the specific example shown in FIG. 2, the modulation symbol at time slot 1 corresponds to pilot data, which may be periodically transmitted to assist the receiver units synchronize and perform

channel estimation. Other techniques for distributing pilot data over time and frequency may also be used. Transmission of the pilot modulation symbol typically occurs at a particular rate, which is usually selected to be fast enough to permit accurate tracking of variations in the communication link.

The time slots not used for pilot transmissions can be used to transmit various types of data. For example, frequency subchannels 1 and 2 may be reserved for the transmission of control and broadcast data to the receiver units. The data on these subchannels is generally intended to be received by all receiver units. However, some of the messages on the control channel may be user specific, and may be encoded accordingly.

Voice data and packet data may be transmitted in the remaining frequency subchannels. For the example shown, subchannel 3 at time slots 2 through 9 is used for voice call 1, subchannel 4 at time slots 2 through 9 is used for voice call 2, subchannel 5 at time slots 5 through 9 is used for voice call 3, and subchannel 6 at time slots 7 through 9 is used for voice call 5.

The remaining available frequency subchannels and time slots may be used for transmissions of traffic data. A particular data transmission may occur over multiple subchannels and/or multiple time slots, and multiple data transmissions may occur within any particular time slot. A data transmission may also occur over non-contiguous time slots.

In the example shown in FIG. 2, data 1 transmission uses frequency subchannels 5 through 16 at time slot 2 and subchannels 7 through 16 at time slot 7, data 2 transmission uses subchannels 5 through 16 at time slots 3 and 4 and subchannels 6 through 16 at time slots 5, data 3 transmission uses subchannels 6 through 16 at time slot 6, data 4 transmission uses subchannels 7 through 16 at time slot 8, data 5 transmission uses subchannels 7 through 11 at time slot 9, and data 6 transmission uses subchannels 12 through 16 at time slot 9. Data 1 through 6 transmissions can represent transmissions of traffic data to one or more receiver units.

To provide the transmission flexibility and achieve high performance and efficiency, each frequency subchannel at each time slot for each transmit antenna may be viewed as an independent unit of transmission (a modulation symbol) that may be used to transmit any type of data such as pilot, signaling, broadcast, voice, traffic data, some other data type, or a combination thereof. Flexibility, performance, and efficiency may further be achieved by allowing for independence among the modulation symbols, as described below. For example, each modulation symbol may be generated from a modulation

scheme (e.g., M-PSK, M-QAM, or some other scheme) that results in the best use of the resource at that particular time, frequency, and space.

MIMO System

In a terrestrial communications system (e.g., a cellular system, a broadcast system, a multi-channel multi-point distribution system (MMDS) system, and others), an RF modulated signal from a transmitter unit may reach the receiver unit via a number of transmission paths. The characteristics of the transmission paths typically vary over time due to a number of factors. If more than one transmit or receive antenna is used, and if the transmission paths between the transmit and receive antennas are linearly independent (i.e., one transmission is not formed as a linear combination of the other transmissions), which is generally true to at least an extent, then the likelihood of correctly receiving the transmitted signal increases as the number of antennas increases. Generally, as the number of transmit and receive antennas increases, diversity increases and performance improves.

A MIMO communication system such as the one shown in FIG. 1 employs antennas at both the transmit and receive ends of the communication link. These transmit and receive antennas may be used to provide various forms of "spatial diversity", including "transmit" diversity and "receive" diversity. Spatial diversity is characterized by the use of multiple transmit antennas and one or more receive antennas. Transmit diversity is characterized by the transmission of data over multiple transmit antennas. Typically, additional processing is performed on the data transmitted from the transmit antennas to achieved the desired diversity. For example, the data transmitted from different transmit antennas may be delayed or reordered in time, coded and interleaved across the available transmit antennas, and so on. Receive diversity is characterized by the reception of the transmitted signals on multiple receive antennas, and diversity is achieved by simply receiving the signals via different signal paths.

Spatial diversity may be used to improve the reliability of the communication link with or without increasing the link capacity. This may be achieved by transmitting or receiving data over multiple paths via multiple antennas. Spatial diversity may be dynamically selected based on the characteristics of the communication link to provide the required performance. For example, higher degree of spatial diversity may be provided for some types of communication (e.g., signaling), for some types of services (e.g., voice), for

some communication link characteristics (e.g., low SNR), or for some other conditions or considerations.

The data may be transmitted from multiple antennas and/or on multiple frequency subchannels to obtain the desired diversity. For example, data may be transmitted on: (1) one subchannel from one antenna, (2) one subchannel (e.g., subchannel 1) from multiple antennas, (3) one subchannel from all N_T antennas, (4) a set of subchannels (e.g., subchannels 1 and 2) from one antenna, (5), a set of subchannels from multiple antennas, (6) a set of subchannels from all N_T antennas, or (7) a set of channels from a set of antennas (e.g., subchannel 1 from antennas 1 and 2 at one time slot, subchannels 1 and 2 from antenna 2 at another time slot, and so on). Thus, any combination of subchannels and antennas may be used to provide antenna and frequency diversity.

In the MIMO communication system, the multi-input multi-output channel can be decomposed into a set of N_C independent spatial subchannels. The number of such spatial subchannels is less than or equal to the lesser of the number of the transmitting antennas and the number of receiving antennas (i.e., $N_C \leq N_T$ and $N_C \leq N_R$). If H is the N_R x N_T matrix that gives the channel response for the N_T transmit antennas and the N_R receive antennas at a specific time, and $\underline{x}$ is the N_T-vector inputs to the channel, then the received signal can be expressed as:

$$\underline{y} = \mathbf{H}\underline{x} + \underline{n} \ ,$$

where $\underline{n}$ is an N_R-vector representing noise plus interference. In one embodiment, the eigenvector decomposition of the Hermitian matrix formed by the product of the channel matrix with its conjugate-transpose can be expressed as:

$$\mathbf{H}^{*}\mathbf{H} = E\Lambda E^{*} \ ,$$

where the symbol "*" denotes conjugate-transpose, E is the eigenvector matrix, and Λ is a diagonal matrix of eigenvalues, both of dimension $N_T x N_T$.

The transmitter converts (i.e., pre-conditions) a set of N_T modulation symbols $\underline{b}$ using the eigenvector matrix E. The transmitted modulation symbols from the N_T transmit antennas can be expressed as:

$$\underline{x} = E\underline{b} \ .$$

For all antennas, the pre-conditioning of the modulation symbols can be achieved by a matrix multiply operation expressed as:

$$\begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_{N_T} \end{bmatrix} = \begin{bmatrix} e_{11}, & e_{12}, & & e_{1N_T} \\ e_{21}, & e_{22}, & & e_{2N_T} \\ & & & \\ e_{N_T1}, & e_{N_T2}, & & e_{N_TN_T} \end{bmatrix} \begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_{N_T} \end{bmatrix} \qquad \text{Eq (1)}$$

where b_1, b_2, ... and b_{NT} are respectively the modulation symbols for a particular frequency subchannel at transmit antennas 1, 2, ... N_T, where each modulation symbol can be generated using, for example, M-PSK, M-QAM, and so on, as described below;

E = is the eigenvector matrix related to the transmission characteristics from transmit antennas to the receive antennas; and

x_1, x_2, ... x_{NT} are the pre-conditioned modulation symbols, which can be expressed as:

$$x_1 = b_1 \cdot e_{11} + b_2 \cdot e_{12} + \ \ldots \ + b_{N_T} \cdot e_{1N_T} \ ,$$

$$x_2 = b_1 \cdot e_{21} + b_2 \cdot e_{22} + \ \ldots \ + b_{N_T} \cdot e_{2N_T} \ , \text{ and}$$

$$x_{N_T} = b_1 \cdot e_{N_T1} + b_2 \cdot e_{N_T2} + \ \ldots \ + b_{N_T} \cdot e_{N_TN_T} \ .$$

The received signal may be expressed as:

$$\underline{\mathbf{y}} = \mathbf{H}E\underline{\mathbf{b}} + \underline{\mathbf{n}} \ .$$

The receiver performs a channel-matched-filter operation, followed by multiplication by the right eigenvectors. The result of the channel-matched-filter operation is the vector $\underline{\mathbf{z}}$, which can be expressed as:

$$\underline{\mathbf{z}} = E^{*}\mathbf{H}^{*}\mathbf{H}E\underline{\mathbf{b}} + E^{*}\mathbf{H}^{*}\underline{\mathbf{n}} = \Lambda\underline{\mathbf{b}} + \underline{\hat{\mathbf{n}}} \ ,$$

where the new noise term has covariance that can be expressed as:

$$E(\underline{\hat{n}}\underline{\hat{n}}^{*}) = E(E^{*}\mathbf{H}^{*}\underline{\mathbf{n}}\underline{\mathbf{n}}^{*}\mathbf{H}E) = E^{*}\mathbf{H}^{*}\mathbf{H}E = \Lambda \ ,$$

i.e., the noise components are independent and have variance given by the eigenvalues. The SNR of the ith component of $\underline{\mathbf{z}}$ is λ_i, the ith diagonal element of Λ.

An embodiment of the MIMO processing is described in further detail below and in U.S Patent Application Serial No. 09/532,491, entitled "HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM EMPLOYING MULTI-CARRIER MODULATION," filed March 22, 2000,

assigned to the assignee of the present application and incorporated herein by reference.

Each of the N_c spatial subchannels in the MIMO channel as described in the above embodiment is also referred to as an eigenmode if these channels are independent of each other. For the MIMO mode, one modulation symbol can be transmitted on each of the eigenmodes in each frequency subchannel. Since the SNR may be different for each eigenmode, the number of bits that may be transmitted over each eigenmode may also be different. As noted above, each eigenmode of each frequency subchannel is also referred to as a transmission channel.

In other embodiments, the spatial subchannels can be created differently. For example, a spatial subchannel can be defined as the transmissions from one transmitter antenna to all of the receiver antennas.

As used herein, the MIMO mode includes full channel state information (full-CSI) and partial-CSI processing modes. For both full-CSI and partial-CSI, additional transmission paths are provided via spatially separable subchannels. Full-CSI processing utilizes eigenmodes, as described above. Partial-CSI processing does not utilize eigenmodes, and may involve providing to the transmitter unit (e.g., via feeding back on the reverse link) the SNR for each transmission channel (i.e., receive diversity port), and coding accordingly based on the received SNR.

A number of formulations may be utilized at the receiver unit to provide the requisite information for partial-CSI, including linear and non-linear forms of zero-forcing, channel correlation matrix inversion (CCMI), and minimum mean square error (MMSE), as is known in the art. For example, the derivation of SNRs for a non-linear zero-forcing (partial-CSI) MIMO case is described by P.W. Wolniansky *et al.* in a paper entitled "V-BLAST: An Architecture for Realizing Very High Data Rates Over the Rich-Scattering Wireless Channel," Proc. IEEE ISSSE-98, Pisa, Italy. Sept. 30, 1998, and incorporated herein by reference. The eigenvalues from a MIMO formulation are related to the SNRs of the eigenmodes for the full-CSI case. Non-MIMO cases can use an assortment of methods, as is known in the art.

Each transmission channel is associated with a SNR that may be known to both the transmitter and receiver. In this case, the modulation and coding parameters of each modulation symbol can be determined based on the SNR of the corresponding transmission channel. This allows for efficient use of the available frequency subchannels and eigenmodes.

Table 1 lists the number of information bits that may be transmitted in each modulation symbol for a particular level of performance (e.g., 1% frame-error rate, or % FER) for various SNR ranges. For each SNR range, Table 1 also lists a particular modulation scheme selected for use with that SNR range, the number of coded bits that may be transmitted for each modulation symbol for the selected modulation scheme, and the coding rate used to obtain the required number of coded bits/modulation symbol given the supported number of information bits/modulation symbol.

Table 1 lists one combination of modulation scheme and coding rate for each SNR range. The supported bit rate for each transmission channel may be achieved using any one of a number of possible combinations of coding rate and modulation scheme. For example, one information bit per symbol may be achieved using (1) a coding rate of 1/2 and QPSK modulation, (2) a coding rate of 1/3 and 8-PSK modulation, (3) a coding rate of 1/4 and 16-QAM, or (4) some other combination of coding rate and modulation scheme. In Table 1, QPSK, 16-QAM, and 64-QAM are used for the listed SNR ranges. Other modulation schemes such as 8-PSK, 32-QAM, 128-QAM, and so on, may also be employed and are within the scope of the invention.

Table 1

SNR Range	# of Information Bits/Symbol	Modulation Symbol	# of Coded Bits/Symbol	Coding Rate
1.5 – 4.4	1	QPSK	2	1/2
4.4 – 6.4	1.5	QPSK	2	3/4
6.4 – 8.35	2	16-QAM	4	1/2
8.35 – 10.4	2.5	16-QAM	4	5/8
10.4 – 12.3	3	16-QAM	4	3/4
12.3 – 14.15	3.5	64-QAM	6	7/12
14.15 – 15.55	4	64-QAM	6	2/3
15.55 – 17.35	4.5	64-QAM	6	3/4
> 17.35	5	64-QAM	6	5/6

For clarity, various aspects of the invention are described for an OFDM system and, in many instances, for an OFDM system operating in a MIMO mode. However, the encoding and processing techniques described herein may generally be applied to various communication systems such as, for example, (1) an OFDM system operating without MIMO, (2) a MIMO system operating without OFDM (i.e., operating based on a single frequency

subchannel, i.e., a single RF carrier, but multiple spatial subchannels), (3) a MIMO system operating with OFDM, and (4) others. OFDM is simply one technique for subdividing a wideband channel into a number of orthogonal frequency subchannels.

<u>Encoding</u>

FIG. 3A is a block diagram of an embodiment of a parallel concatenated convolutional encoder 114x, which is often referred to as a Turbo encoder. Turbo encoder 114x represents one implementation of the forward error correction (FEC) portion of encoder 114 in FIG. 1 and may be used to encode data for transmission over one or more transmission channels.

The encoding within encoder 114 may include error correction coding or error detection coding, or both, which are used to increase the reliability of the link. The encoding may include, for example, cyclic redundancy check (CRC) coding, convolutional coding, Turbo coding, Trellis coding, block coding (e.g., Reed-Solomon coding), other types of coding, or a combination thereof. For a wireless communication system, a packet of data may be initially encoded with a particular CRC code, and the CRC bits are appended to the data packet. Additional overhead bits may also be appended to the data packet to form a formatted data packet, which is then encoded with a convolutional or Turbo code. As used herein, "information bits" refer to bits provided to the convolutional or Turbo encoder, including transmitted data bits and bits used to provide error detection or correction capability for the transmitted bits.

As shown in FIG. 3A, Turbo encoder 114x includes two constituent encoders 312a and 312b, and a code interleaver 314. Constituent encoder 312a receives and encodes the information bits, x, in accordance with a first constituent code to generate a first sequence of tail and parity bits, y. Code interleaver 314 receives and interleaves the information bits in accordance with a particular interleaving scheme. Constituent encoder 312b receives and encodes the interleaved bits in accordance with a second constituent code to generate a second sequence of tail and parity bits, z. The information bits, tail bits, and parity bits from encoders 312a and 312b are provided to the next processing element (channel interleaver 116).

FIG. 3B is a diagram of an embodiment of a Turbo encoder 114y, which is one implementation of Turbo encoder 114x and may also be used within encoder 114 in FIG. 1. In this example, Turbo encoder 114y is a rate 1/3 encoder that provides two parity bits, y and z, for each information bit x.

In the embodiment shown in FIG. 3B, each constituent encoder 322 of Turbo encoder 114y implements the following transfer function for the constituent code:

$$G(D) = \left[1 \quad \frac{n(D)}{d(D)} \right] ,$$

where

$$n(D) = 1 + D + D^3 \qquad , \textit{and}$$

$$d(D) = 1 + D^2 + D^3$$

Other constituent codes may also be used and are within the scope of the invention.

Each constituent encoder 322 includes a number of series coupled delay elements 332, a number of modulo-2 adders 334, and a switch 336. Initially, the states of delay elements 332 are set to zeros and switch 336 is in the up position. Then, for each information bit in a data packet, adder 334a performs modulo-2 addition of the information bit with the output bit from adder 334c and provides the result to delay element 332a. Adder 334b receives and performs modulo-2 addition of the bits from adder 334a and delay elements 332a and 332c, and provides the parity bit y. Adder 334c performs modulo-2 addition of the bits from delay elements 332b and 332c.

After all N information bits in the data packet have been encoded, switch 336 is moved to the down position and three zero ("0") bits are provided to the constituent encoder 322a. Constituent encoder 322a then encodes the three zero bits and provides three tail systematic bits and three tail parity bits.

For each packet of N information bits, constituent encoder 322a provides N information bits x, the first three tail systematic bits, N parity bits y, and the first three tail parity bits, and constituent encoder 322b provides the second three tail systematic bits, N parity bits z, and the last three tail parity bits. For each packet, encoder 114y provides N information bits, six tail systematic bits, N+3 parity bits from encoder 322a, and N+3 parity bits from encoder 322b.

Code interleaver 314 may implement any one of a number of interleaving schemes. In one specific interleaving scheme, the N information bits in the packet are written, by row, into a 2^5-row by 2^n-column array, where n is the smallest integer such that $N \leq 2^{5+n}$. The rows are then shuffled according to a bit-reversal rule. For example, row 1 ("00001") is swapped with row 16 ("10000"), row 3 ("00011") is swapped with row 24 ("11000"), and so on. The bits

within each row are then permutated (i.e., rearranged) according to a row-specific linear congruential sequence (LCS). The LCS for row k may be defined as $x_k(i+1) = \{x_k(i) + c_k\} \bmod 2^n$, where $i = 0, 1, \ldots 2^n-1$, $x_k(0) = c_k$, and c_k is a specific value selected for each row and is further dependent on the value for n. For permutation in each row, the i^{th} bit in the row is placed in location x(i). The bits in code interleaver 314 are then read out by column.

The above LCS code interleaving scheme is described in further detail in commonly assigned U.S. Patent Application Serial No. 09/205,511, entitled "TURBO CODE INTERLEAVER USING LINEAR CONGRUENTIAL SEQUENCES," filed December 4, 1998, and in a document entitled "C.S0002-A-1 Physical Layer Standard for cdma2000 Spread Spectrum Systems" (hereinafter referred to as the cdma2000 standard), both of which are incorporated herein by reference.

Other code interleaver may also be used and are within the scope of the invention. For example, a random interleaver or a symmetrical-random (S-random) interleaver may also be used instead of the linear congruential sequence interleaver described above.

For clarity, the data coding is specifically described based on a Turbo code. Other coding schemes may also be used and are within the scope of the invention. For example, the data may be coded with a convolutional code, a block code, a concatenated code comprised of a combination of block, convolutional, and/or Turbo codes, or some other code. The data may be coded in accordance with a "base" code, and the coded bits may thereafter be processed (e.g., punctured) based on the capabilities of the transmission channels used to transmit the data.

Channel Interleaving

Referring back to FIG. 1, the coded bits from encoder 114 are interleaved by channel interleaver 116 to provide temporal and frequency diversity against deleterious path effects (e.g., fading). Moreover, since coded bits are subsequently grouped together to form non-binary symbols that are then mapped to modulation symbols, the interleaving further ensures that the coded bits that form each modulation symbol are not located close to each other (temporally). For static additive white Gaussian noise (AWGN) channels, the channel interleaving is less critical when a Turbo encoder is also employed, since the code interleaver effectively performs similar functions.

Various interleaving schemes may be used for the channel interleaver. In one interleaving scheme, the coded bits (i.e., the information, tail, and parity

bits) for each packet are written (linearly) to rows of memory. The bits in each row may then be permutated (i.e., rearranged) based on (1) a bit-reversal rule, (2) a linear congruential sequence (such as the one described above for the code interleaver), (3) a randomly generated pattern, (4) or a permutation pattern generated in some other manner. The rows are also permutated in accordance with a particular row permutation pattern. The permutated coded bits are then retrieved from each column and provided to puncturer 117.

In an embodiment, the channel interleaving is performed individually for each bit stream in a packet. For each packet, the information bits x, the tail and parity bits y from the first constituent encoder, and the tail and parity bits z from the second constituent encoder may be interleaved by three separate interleavers, which may employ the same or different channel interleaving schemes. This separate interleaving allows for flexible puncturing on the individual bit streams.

The interleaving interval may be selected to provide the desired temporal and frequency diversity. For example, coded bits for a particular time period (e.g., 10 msec, 20 msec, or some other) and/or for a particular number of transmission channels may be interleaved.

<u>Puncturing</u>

As noted above, for an OFDM communication system, the number of information bits that may be transmitted for each modulation symbol is dependent on the SNR of the transmission channel used to transmit the modulation symbol. And for an OFDM system operated in the MIMO mode, the number of information bits that may be transmitted for each modulation symbol is dependent on the SNR of the frequency subchannel and spatial subchannel used to transmit the modulation symbol.

In accordance with an aspect of the invention, a number of coding/puncturing schemes may be used to generate the coded bits (i.e., information, tail, and parity bits) for transmission. In a first coding/puncturing scheme, a particular base code and common puncturing is applied for all transmission channels. In a second coding/puncturing scheme, the same base code but variable puncturing is applied for the transmission channels. The variable puncturing is dependent on the SNR of the transmission channels.

FIG. 4A is a flow diagram of an embodiment for generating the required coded bits for a data transmission, which employs the base code and common puncturing scheme. Initially, the SNR for each transmission channel (i.e., each eigenmode of each frequency subchannel) is determined, at step 412. For an

OFDM system not operated in the MIMO mode, only one eigenmode is supported and thus only one SNR is determined for each frequency subchannel. The SNR for each transmission channel may be determined based on the transmitted pilot reference or via some other mechanism.

At step 414, the number of information bits per modulation symbol supported by each transmission channel is determined based on its SNR. A table that associates a range of SNR with each specific number of information bits/modulation symbol, such as Table 1, may be used. However, finer quantization than the 0.5-bit step size for the information bits shown in Table 1 may be used. A modulation scheme is then selected for each transmission channel such that the number of information bits/modulation symbol can be transmitted, at step 416. The modulation scheme may also be selected to take into account other factors (e.g., coding complexity), as described in further detail below.

At step 418, the total number of information bits that may be transmitted in each time slot for all transmission channels is determined. This can be achieved by summing the number of information bits/modulation symbol determined for all transmission channels. Similarly, the total number of coded bits that may be transmitted in each time slot for all transmission channels is determined, at step 420. This can be achieved by determining the number of coded bits/modulation symbol for each modulation scheme selected in step 416, and summing the number of coded bits for all transmission channels.

At step 422, the total number of information bits determined in step 418 is encoded with a particular encoder. If a Turbo encoder is used, the tail bits and parity bits generated by the encoder are punctured to obtain the total number of coded bits determined in step 420. The unpunctured coded bits are then grouped into non-binary symbols, which are then mapped to modulation symbols for the transmission channels, at step 426.

The first coding/puncturing scheme is relatively simple to implement since the same base code and puncturing scheme are used for all transmission channels. The modulation symbol for each transmission channel represents a point in a signal constellation corresponding to the modulation scheme selected for that transmission channel. If the distribution of the SNR for the transmission channels is widespread, the distance between the constellation points relative to the noise variance for different signal constellations will vary widely. This may then impact the performance of the system.

FIG. 4B is a flow diagram of an embodiment for generating the required coded bits for a data transmission, which employs the same base code but

variable puncturing scheme. Initially, the SNR for each transmission channel is determined, at step 432. In an embodiment, transmission channels with insufficient SNR are omitted from use for data transmission (i.e., no data is transmitted on poor transmission channels). The number of information bits per modulation symbol supported by each transmission channel is then determined based on its SNR, at step 434. A modulation scheme is next selected for each transmission channel such that the number of information bits/modulation symbol can be transmitted, at step 436. Steps 432, 434, and 436 in FIG. 4B correspond to steps 412, 414, and 416 in FIG. 4A.

At step 438, the transmission channels belonging to the same SNR range are grouped into a segment. Alternatively, ranges can be defined for the number of information bits per modulation symbol (e.g., range 1 covering 1.0 to 1.5 information bits/modulation symbol, range 2 covering 1.5 to 2.0 information bits/modulation symbol, and so on). In this case, transmission channels having number of information bits per modulation symbol within the same range are grouped into a segment.

Each segment includes K_i transmission channels, where K_i can be any integer one or greater. The total number of information bits and total number of coded bits that can be transmitted in each segment are then determined, at step 440. For example, segment i may include K_i transmission channels, each of which may support transmission of N_i information bits/modulation symbol and P_i tail and parity bits/modulation symbol. For each time slot, the total number of information bits that may be transmitted in segment i can be computed as $K_i \cdot N_i$, the total number of tail and parity bits that may be transmitted can be computed as $K_i \cdot P_i$, and the total number of coded bits may be computed as $K_i(N_i + P_i)$.

At step 442, the information bits to be transmitted in each time slot for all segments, which may be computed as $\sum_i K_i N_i$, are encoded with a particular encoder (e.g., a rate 1/3 Turbo encoder such at the one shown in FIG. 3B). At step 444, N_i information bits and N_i/R parity and tail bits are assigned to each transmission channel of segment i, where R is the coding rate of the encoder. The N_i/R parity and tail bits are then punctured to obtain the P_i parity and tail bits required for each transmission channel of the segment, at step 446. At step 448, the N_i information bits and the P_i parity and tail bits for each transmission channel of segment i are mapped to a modulation symbol for the transmission channel.

The second coding/puncturing scheme may provide improved performance over the first scheme, especially if the distribution of SNR for the transmission channels is widespread. Since different modulation schemes and coding rate may be used for different transmission channels, the number of bits transmitted on each transmission channel is typically communicated from the receiver to the transmitter on the reverse link.

Table 1 shows the quantization of the number of information bits/modulation symbol using 0.5-bit step size. The quantization granularity may be reduced (i.e., to be finer than 0.5-bit) if each segment (and not each transmission channel) is required to support an integer number of information bits. If $K_i \cdot N_i$ is required to be an integer, a larger integer value for K_i allows for a smaller step size for N_i. The quantization granularity may be further reduced if the quantization is allowed to be carried from segment to segment. For example, if one bit needs to be rounded-off in one segment, one bit may be rounded-up in the next segment, if appropriate. The quantization granularity may also be reduced if the quantization is allowed to be carried over multiple time slots.

To support an OFDM system (especially one operated in the MIMO mode) whereby different SNR may be achieved for the transmission channels, a flexible puncturing scheme may be used in conjunction with a common base encoder (e.g., a rate 1/3 Turbo encoder) to achieve the necessary coding rates. This flexible puncturing scheme may be used to provide the necessary number of tail and parity bits for each segment. For a high coding rate in which more tail and parity bits are punctured than retained, the puncturing may be efficiently achieved by retaining the required number of tail and parity bits as they are generated by the encoder and discarding the others.

As an example, a segment may include 20 16-QAM modulation symbols and has a SNR that supports transmission of 2.75 information bits/modulation symbol. For this segment, 55 information bits (55 = 20x2.75) may be transmitted in 20 modulation symbols. Each 16-QAM modulation symbol is formed with four coded bits, and 80 coded bits are needed for 20 modulation symbols. The 55 information bits may be encoded with a rate 1/3 encoder to generate 122 tail and parity bits and 55 information bits. These 122 tail and parity bits may be punctured to provide the 35 tail and parity bits required for the segment, which in combination with the 55 information bits comprise the 80 coded bits.

Referring back to FIG. 1, puncturer 117 receives the interleaved information and parity bits from channel interleaver 116, punctures (i.e.,

deletes) some of the tail and parity bits to achieve the desired coding rate(s), and multiplexes the unpunctured information, tail, and parity bits into a sequence of coded bits. The information bits (which are also referred to as systematic bits) may also be punctured along with the tail and parity bits, and this is within the scope of the invention.

FIG. 3C is a diagram of an embodiment of a puncturer 117x, which may be used to provide variable puncturing of coded bits. Puncturer 117x is one implementation of puncturer 117 in FIG. 1. Using a set of counters, puncturer 117x performs puncturing to retain P_i tail and parity bits out of Q_i tail and parity bits generated by the encoder for segment i.

Within puncturer 117x, the interleaved tail and parity bits y_{INT} and z_{INT} from the two constituent encoders of the Turbo encoder are provided to two inputs of a switch 342. Switch 342 provides either the y_{INT} tail and parity bits or the z_{INT} tail and parity bits to line 343, depending on a control signal from a toggle unit 348. Switch 342 ensures that the tail and parity bits from the two constituent encoders are evenly selected by alternating between the two tail and parity bit streams.

A first counter 352 performs modulo-Q addition and wraps around after its content reaches beyond Q-1. A second counter 354 counts (by one) the Q tail and parity bits. For each segment, both counters 352 and 354 are initially set to zero, switch 342 is in the up position, and the first tail or parity bit $y_{INT,0}$ is provided from multiplexer 346 by closing a switch 344 and appropriately controlling the multiplexer. For each subsequent clock cycle, counter 352 is incremented by P and counter 354 is incremented by one. The value of counter 352 is provided to a decision unit 356. If counter 352 experiences a modulo-Q operation (i.e., the content of counter 352 wraps around), the tail or parity bit on line 343 is provided through switch 344 to multiplexer 346, which then provides the tail or parity bit as an output coded bit. Each time a tail or parity bit is provided from multiplexer 346, toggle unit 348 toggles the state of the control signal, and the other tail and parity bit stream is provided to line 343. The process continues until all Q_i tail and parity bits in the segment are exhausted, as indicated by comparison unit 358.

Other puncturing patterns may also be used and are within the scope of the invention. To provide good performance, the number of tail and parity bits to be punctured should be balanced between the two constituent codes (i.e., approximately equal number of y_{INT} and z_{INT} tail and parity bits are selected) and the unpunctured bits should be distributed relatively evenly over the code block for each segment.

In certain instances, the number of information bits may be less than the capacity of the transmission channels. In such instances, the available and unfilled bit positions may be filled with zero padding, by repeating some of the coded bits, or by some other scheme. The transmit power may also be reduced for some schemes.

Gray Mapping

In an embodiment, for each modulation scheme (e.g., QPSK, 16-QAM, 64-QAM, and so on) selected for use, the points in the signal constellation for the modulation scheme are defined using Gray mapping. The Gray mapping reduces the number of bit errors for more likely error events, as described in further detail below.

FIG. 5 is a diagram of a signal constellation for 16-QAM and a specific Gray mapping scheme. The signal constellation for 16-QAM includes 16 points, each of which is associated with a specific 4-bit value. For Gray mapping, the 4-bit values are associated with the points in the signal constellation such that the values for adjacent points (in the horizontal or vertical direction) differ by only one bit position. The values for points further way differ by more bit positions (e.g., the values for adjacent points in the diagonal direction differ by two bit positions).

Each group of four coded bits $(b_1 b_2 b_3 b_4)$ is mapped to a specific point in the signal constellation associated with the same value as that of the four coded bits. For example, a value of ("0111") for the four coded bits is mapped to a point 512 in the signal constellation. This point then represents the modulation symbol for the four coded bits. For 16-QAM, each modulation symbol represents a specific one of the 16 points in the signal constellation, with the specific point being determined by the value of the four coded bits. Each modulation symbol can be expressed as a complex number $(c + jd)$ and provided to the next processing element (i.e., MIMO processor 120 in FIG. 1).

At the receiver unit, the modulation symbols are received in the presence of noise and typically do not map to the exact location in the signal constellation. For the above example, the received modulation symbol for the transmitted coded bits ("0111") may not map to point 512 at the receiver unit. The noise may have caused the received modulation symbol to be mapped to another location in the signal constellation. Typically, there is greater likelihood of the received modulation symbol being mapped to a location near the correct location (e.g., near the points for "0101", "0011", "0110", or "1111"). Thus, the more likely error event is a received modulation symbol being

erroneously mapped to a point adjacent to the correct point. And since adjacent points in the signal constellation have values that differ by only one bit position, the Gray mapping reduces the number of error bits for more likely error events.

FIG. 5 shows a specific Gray mapping scheme for the 16-QAM signal constellation. Other Gray mapping schemes may also be used and are within the scope of the invention. The signal constellations for other modulation schemes (e.g., 8-PSK, 64-QAM, and so on) may also be mapped with similar or other Gray mapping schemes. For some modulation schemes such as 32-QAM and 128-QAM, a partial Gray mapping scheme may be used if a full Gray mapping scheme is not possible. Also, mapping schemes not based on Gray mapping may also be used and are within the scope of the invention.

MIMO Processing

FIG. 6 is a block diagram of an embodiment of a MIMO processor 120x, which is one implementation of MIMO processor 120 in FIG. 1. The modulation symbols may be transmitted on multiple frequency subchannels and possibly from multiple transmit antennas. When operating in the MIMO mode, the transmission on each frequency subchannel and from each transmit antenna represents non-duplicated data.

Within MIMO processor 120x, a demultiplexer (DEMUX) 610 receives and demultiplexes the modulation symbols into a number of subchannel symbol streams, S_1 through S_L, one subchannel symbol stream for each frequency subchannel used to transmit the symbols. Each subchannel symbol stream is then provided to a respective subchannel MIMO processor 612.

Each subchannel MIMO processor 612 may further demultiplex the received subchannel symbol stream into a number of (up to N_T) symbol sub-streams, one symbol sub-stream for each antenna used to transmit the modulation symbols. When the OFDM system is operated in the MIMO mode, each subchannel MIMO processors 612 pre-conditions the (up to) N_T modulation symbols in accordance with equation (1) described above to generate pre-conditioned modulation symbols, which are subsequently transmitted. In the MIMO mode, each pre-conditioned modulation symbol for a particular frequency subchannel of a particular transmit antenna represents a linear combination of (weighted) modulation symbols for up to N_T transmit antennas. Each of the (up to) N_T modulation symbols used to generate each pre-conditioned modulation symbol may be associated with a different signal constellation.

For each time slot, (up to) N_T pre-conditioned modulation symbols may be generated by each subchannel MIMO processor 612 and provided to (up to) N_T symbol combiners 616a through 616t. For example, subchannel MIMO processor 614a assigned to frequency subchannel 1 may provide up to N_T pre-conditioned modulation symbols for frequency subchannel 1 of antennas 1 through N_T. Similarly, subchannel MIMO processor 612*l* assigned to frequency subchannel L may provide up to N_T symbols for frequency subchannel L of antennas 1 through N_T. Each combiner 616 receives the pre-conditioned modulation symbols for the L frequency subchannels, combines the symbols for each time slot into a modulation symbol vector, V, and provides the modulation symbol vector to the next processing stage (i.e., modulator 122).

MIMO processor 120x thus receives and processes the modulation symbols to provide N_T modulation symbol vectors, V_1 through V_T, one modulation symbol vector for each transmit antenna. The collection of L pre-conditioned modulation symbols for each time slot of each antenna form a modulation symbol vector V of dimensionality L. Each element of the modulation symbol vector V is associated with a specific frequency subchannel having a unique subcarrier on which the modulation symbol is conveyed. The collection of the L modulation symbols are all orthogonal to one another. If not operating in a "pure" MIMO mode, some of the modulation symbol vectors may have duplicate information on specific frequency subchannels for different transmit antennas.

Subchannel MIMO processor 612 may be designed to provide the necessary processing to implement full channel state information (full-CSI) or partial-CSI processing for the MIMO mode. Full CSI includes sufficient characterization of the propagation path (i.e., amplitude and phase) between all pairs of transmit and receive antennas for each frequency subchannel. Partial CSI may include, for example, the SNR of the spatial subchannels. The CSI processing may be performed based on the available CSI information and on the selected frequency subchannels, transmit antennas, and so on. The CSI processing may also be enabled and disabled selectively and dynamically. For example, the CSI processing may be enabled for a particular data transmission and disabled for some other data transmissions. The CSI processing may be enabled under certain conditions, for example, when the communication link has adequate SNR. Full-CSI and partial-CSI processing is described in further detail in the aforementioned U.S Patent Application Serial No. 09/532,491.

FIG. 6 also shows an embodiment of modulator 122. The modulation symbol vectors V_1 through V_T from MIMO processor 120x are provided to

modulators 114a through 114t, respectively. In the embodiment shown in FIG. 6, each modulator 114 includes an IFFT 620, cycle prefix generator 622, and an upconverter 624.

IFFT 620 converts each received modulation symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using the inverse fast Fourier transform (IFFT). IFFT 620 can be designed to perform the IFFT on any number of frequency subchannels (e.g., 8, 16, 32, and so on). In an embodiment, for each modulation symbol vector converted to an OFDM symbol, cycle prefix generator 622 repeats a portion of the time-domain representation of the OFDM symbol to form a transmission symbol for the specific antenna. The cyclic prefix insures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects. The implementation of IFFT 620 and cycle prefix generator 622 is known in the art and not described in detail herein.

The time-domain representations from each cycle prefix generator 622 (i.e., the "transmission" symbols for each antenna) are then processed by upconverter 624, converted into an analog signal, modulated to a RF frequency, and conditioned (e.g., amplified and filtered) to generate an RF modulated signal, which is then transmitted from the respective antenna 124.

OFDM modulation is described in further detail in a paper entitled "Multicarrier Modulation for Data Transmission : An Idea Whose Time Has Come," by John A.C. Bingham, IEEE Communications Magazine, May 1990, which is incorporated herein by reference.

For an OFDM system not operated in the MIMO mode, MIMO processor 120 may be removed or disabled and the modulation symbols may be grouped into the modulation symbol vector V without any pre-conditioning. This vector is then provided to modulator 122. And for an OFDM system operated with transmit diversity (and not in the MIMO mode), demultiplexer 614 may be removed or disabled and the (same) pre-conditioned modulation symbols are provided to (up to) N_T combiners.

As shown in FIG. 2, a number of different transmissions (e.g., voice, signaling, data, pilot, and so on) may be transmitted by the system. Each of these transmissions may require different processing.

FIG. 7 is a block diagram of an embodiment of a system 110y capable of providing different processing for different transmissions. The aggregate input data, which includes all information bits to be transmitted by system 110y, is provided to a demultiplexer 710. Demultiplexer 710 demultiplexes the input

data into a number of (K) channel data streams, B_1 through B_K. Each channel data stream may correspond to, for example, a signaling channel, a broadcast channel, a voice call, or a traffic data transmission. Each channel data stream is provided to a respective encoder/channel interleaver/puncturer/symbol mapping element 712 that encodes the data using a particular encoding scheme selected for that channel data stream, interleaves the encoded data based on a particular interleaving scheme, punctures the interleaved code bits, and maps the interleaved data into modulation symbols for the one or more transmission channels used for transmitting that channel data stream.

The encoding can be performed on a per channel basis (i.e., on each channel data stream, as shown in FIG. 7). However, the encoding may also be performed on the aggregate input data (as shown in FIG. 1), on a number of channel data streams, on a portion of a channel data stream, across a set of frequency subchannels, across a set of spatial subchannels, across a set of frequency subchannels and spatial subchannels, across each frequency subchannel, on each modulation symbol, or on some other unit of time, space, and frequency.

The modulation symbol stream from each encoder/channel interleaver/puncturer/symbol mapping element 712 may be transmitted on one or more frequency subchannels and via one or more spatial subchannels of each frequency subchannel. A MIMO processor 120y receives the modulation symbol streams from elements 712. Depending on the mode to be used for each modulation symbol stream, MIMO processor 120y may demultiplex the modulation symbol stream into a number of subchannel symbol streams. In the embodiment shown in FIG. 7, modulation symbol stream S_1 is transmitted on one frequency subchannel and modulation symbol stream S_K is transmitted on L frequency subchannels. The modulation stream for each frequency subchannel is processed by a respective subchannel MIMO processor, demultiplexed, and combined in similar manner as that described in FIG. 6 to form a modulation symbol vector for each transmit antenna.

In general, the transmitter unit codes and modulates data for each transmission channel based on information descriptive of the channel's transmission capability. This information is typically in the form of partial-CSI or full-CSI described above. The partial or full-CSI for the transmission channels to be used for a data transmission is typically determined at the receiver unit and reported back to the transmitter unit, which then uses the information to code and modulate data accordingly. The techniques described herein are applicable for multiple parallel transmission channels supported by

MIMO, OFDM, or any other communication scheme (e.g., a CDMA scheme) capable of supporting multiple parallel transmission channels.

<u>Demodulation and Decoding</u>

FIG. 8 is a block diagram of an embodiment of a decoding portion of system 150. For this embodiment, a Turbo encoder is used to encode the data prior to transmission. A Turbo decoder is correspondingly used to decode the received modulation symbols.

As shown in FIG. 8, the received modulation symbols are provided to a bit log-likelihood ratio (LLR) calculation unit 158x, which calculates the LLRs of the bits that make up each modulation symbol. Since a Turbo decoder operates on LLRs (as oppose to bits), bit LLR calculation unit 158x provides an LLR for each received coded bit. The LLR for each received coded bit is the logarithm of the probability that the received coded bit is a zero divided by the probability that the received coded bit is a one.

As described above, M coded bits (b_1, b_2, ... b_M) are grouped to form a single non-binary symbol S, which is then mapped to a modulation symbol T(S) (i.e., modulated to a high-order signal constellation). The modulation symbol is processed, transmitted, received, and further processed to provide a received modulation symbol R(S). The LLR of coded bit b_m in the received modulation symbol can be computed as:

$$\begin{aligned} LLR(b_m) &= \log\left(\frac{P(b_m = 0)}{P(b_m = 1)}\right) \\ &= \log\left(P\left(R(S) \mid b_m = 0\right)\right) - \log\left(P\left(R(S) \mid b_m = 1\right)\right) \qquad \text{Eq (2)} \\ &= \log\left(\sum_{T(S):b_m=0} P\left(R(S) \mid T(S)\right)\right) - \log\left(\sum_{T(S):b_m=1} P\left(R(S) \mid T(S)\right)\right) \end{aligned}$$

where $P\left(R(S) \mid b_m = 0\right)$ is the probability of bit b_m being a zero based on the received symbol R(S). Approximations may also be used in computing the LLRs.

De-puncturer 159 then inserts "erasures" for code bits that have been deleted (i.e., punctured) at the transmitter. The erasures typically have a value of zero ("0"), which is indicative of the punctured bit being equally likely to be a zero or a one.

From equation (2), it can be noted that the LLRs for the received coded bits within a modulation symbol tend to be correlated. This correlation can be

broken up by interleaving the coded bits prior to modulation. As shown in FIG. 1, the channel interleaving advantageously performs the decorrelation of the coded bits in each modulation symbol.

The coded bit LLRs are provided to a channel deinterleaver 160 and deinterleaved in a manner complementary to the channel interleaving performed at the transmitter. The channel deinterleaved LLRs corresponding to the received information, tail, and parity bits are then provided to a Turbo decoder 162x.

Turbo decoder 162x includes summers 810a and 810b, decoders 812a and 812b, a code interleaver 814, a code deinterleaver 816, and a detector 818. In an embodiment, each decoder 812 is implemented as a soft-input/soft-output (SISO) maximum a posterior (MAP) decoder.

Summer 810a receives and sums the LLRs of the received information bits, LLR(x'), and the extrinsic information from deinterleaver 816 (which is set to zeros on the first iteration), and provides refined LLRs. The refined LLRs are associated with greater confidence in the detected values of the received information bits.

Decoder 812a receives the refined LLRs from summer 810a and the LLRs of the received tail and parity bits from the first constituent encoder, LLR(y'), and decodes the received LLRs to generate extrinsic information indicative of corrections in the probability values for the received information bits. The extrinsic information from decoder 812a are summed with the received information bit LLRs by summer 810b, and the refined LLRs are stored to code interleaver 814. Code interleaver 814 implements the same code interleaving used at the Turbo encoder (e.g., the same as code interleaver 314 in FIG. 3B).

Decoder 812b receives the interleaved LLRs from interleaver 814 and the LLRs of the received tail and parity bits from the second constituent encoder, LLR(z'), and decodes the received LLRs to generate extrinsic information indicative of further corrections in the probability values for the received information bits. The extrinsic information from decoder 812b is stored to code deinterleaver 816, which implements a deinterleaving scheme complementary to the interleaving scheme used for interleaver 814.

The decoding of the received coded bit LLRs is iterated a number of times. With each iteration, greater confidence is gained for the refined LLRs. After all the decoding iterations have been completed, the final refined LLRs are provided to detector 818, which provides values for the received information bits based on the LLRs.

Other types of decoder may also be used beside the SISO MAP decoder such as one that implements the soft output Viterbi algorithm (SOVA). The design of the decoder is typically dependent on the particular Turbo coding scheme used at the transmitter.

Turbo decoding is described in greater detail by Steven S. Pietrobon in a paper entitled "Implementation and Performance of a Turbo/Map Decoder," International Journal of Satellite Communications, Vol. 16, 1998, pp. 23-46, which is incorporated herein by reference.

Modulation Scheme and Coding Rate

The achieved SNR of each transmission channel supports a particular number of information bits per modulation symbol (i.e., a particular information bit rate) for a desired level of performance (e.g., 1% FER). This information bit rate may be supported by a number of different modulation schemes. For example, a bit rate of 1.5 information bits/modulation symbol may be supported by QPSK, 8-PSK, 16-QAM, or any higher order modulation scheme. Each modulation scheme is able to transmit a particular number of coded bits per modulation symbol.

Depending on the selected modulation scheme, a corresponding coding rate is selected such that the required number of coded bits is provided for the number of information bits for each modulation symbol. For the above example, QPSK, 8-PSK, and 16-QAM are respectively able to transmit 2, 3, and 4 coded bits per modulation symbol. For an information bit rate of 1.5 information bits/modulation symbol, coding rates of 3/4, 1/2, and 3/8 are used to generate the required number of coded bits for QPSK, 8-PSK, and 16-QAM, respectively. Thus, different combinations of modulation scheme and coding rate may be used to support a particular information bit rate.

In certain embodiments of the invention, a "weak" binary code (i.e., a high coding rate) is used in conjunction with a low-order modulation scheme for the supported bit rate. Through a series of simulation, it is observed that a lower order modulation scheme in combination with a weaker code may offer better performance than a higher order modulation scheme with a stronger code. This result may be explained as follows. The LLR decoding metrics of binary Turbo codes in an AWGN channel is near optimal for the Turbo decoding algorithm. However, for the Gray mapped high-order modulation scheme, the optimal LLR metrics are generated for each received modulation symbol and not each received bit. The symbol LLR metrics are then broken to yield bit LLR metrics for the binary code decoder. Some information is lost

during the break-up process, and using the bit decoding metrics may result in non-optimal performance. The lower order modulation schemes correspond to fewer bits per symbol, which may experience less of the break-up loss and therefore provide better performance than the higher order modulation scheme counterparts.

In accordance with an aspect of the invention, in order to achieve certain spectrum efficiency, a code with a coding rate of between, and inclusive of, $n/(n+1)$ to $n/(n+2)$ is used with an appropriate modulation scheme, where n is the number of information bits per modulation symbol. This coding rate may be easily achieved with a fixed code (e.g., the rate 1/3 Turbo code described above) in combination with a variable puncturing scheme. To achieve a high coding rate, the tail and parity bits may be heavily punctured and the unpunctured tail and parity bits may be evenly distributed over the information bits.

Framing

For many communication systems, it is convenient to define data packets (i.e., logical frames) with fixed sizes. For example, a system may define three different packets having sizes of 1024, 2048, and 4096 bits. These defined data packets simplify some of the processing at both the transmitter and receiver.

For an OFDM system, a physical frame may be defined to include (1) an integer number of OFDM symbols, (2) a particular number of modulation symbols on one or more transmission channels, (3) or some other units. As described above, because of the time-variant nature of the communication link, the SNR of the transmission channels may vary over time. Consequently, the number of information bits which may be transmitted on each time slot for each transmission channel will likely vary over time, and the number of information bits in each physical frame will also likely vary over time.

In one embodiment, a logical frame is defined such that it is independent of the OFDM symbols. In this embodiment, the information bits for each logical frame are encoded/punctured, and the coded bits for the logical frame are grouped and mapped to modulation symbols. In one simple implementation, the transmission channels are sequentially numbered. The coded bits are then used to form as many modulation symbols as needed, in the sequential order of the transmission channels. A logical frame (i.e., a data packet) may be defined to start and end at modulation symbol boundaries. In this implementation, the logical frame may span more than one OFDM symbol

and may further cross OFDM symbol boundaries. Moreover, each OFDM symbol may include coded bits from multiple data packets.

In another embodiment, a logical frame is defined based on a physical unit. For example, a logical frame may be defined to include (1) a number of modulation symbols on one or more transmission channels, (2) one or more OFDM symbols, or (3) a number of modulation symbols defined in some other manner.

The use of punctured binary Turbo code and Gray mapping (BTC-GM) for high-order modulation provides numerous advantages. The BTC-GM scheme is simpler to implement than the more optimal but more complicated Turbo trellis coded modulation (TTCM) scheme, yet can achieve performance close to that of TTCM. The BTC-GM scheme also provides a high degree of flexibility because of the ease of implementing different coding rate by simply adjusting the variable puncturing. The BTC-GM scheme also provides robust performance under different puncturing parameters. Also, currently available binary Turbo decoders may be used, which may simply the implementation of the receiver. However, in certain embodiments, other coding schemes may also be used and are within the scope of the invention.

The foregoing description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. In a wireless communication system, a method for preparing data for transmission on a plurality of transmission channels, wherein each transmission channel is operative to transmit a respective sequence of modulation symbols, the method comprising:

determining a number of information bits per modulation symbol supported by each transmission channel;

identifying a modulation scheme for each transmission channel such that the determined number of information bits per modulation symbol is supported;

determining a coding rate for each transmission channel based at least on the determined number of information bits per modulation symbol and the identified modulation scheme for the transmission channel, wherein at least two transmission channels are associated with different coding rates;

encoding a plurality of information bits in accordance with a particular encoding scheme to provide a plurality of coded bits;

puncturing the plurality of coded bits in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the plurality of transmission channels; and

adjusting the puncturing to achieve the different coding rates for the at least two transmission channels.

2. The method of claim 1, wherein the wireless communication system is a multiple-input multiple-output (MIMO) system with a plurality of transmit antennas and a plurality of receive antennas.

3. The method of claim 1, wherein the wireless communication system is an orthogonal frequency division modulation (OFDM) communication system.

4. The method of claim 3, wherein the OFDM communication system is operated as a multiple-input multiple-output (MIMO) system with a plurality of transmit antennas and a plurality of receive antennas

5. The method of claim 4, wherein the OFDM system is operative to transmit data on a plurality of frequency subchannels, and wherein each transmission channel corresponds to a spatial subchannel of a frequency subchannel in the OFDM system.

6. The method of claim 1, wherein the puncturing is based on transmission capabilities of the plurality of transmission channels.

7. The method of claim 6, wherein the transmission capabilities are determined from channel state information (CSI) derived for the plurality of transmission channels.

8. The method of claim 7, wherein the CSI includes signal-to-noise ratio (SNR) information for the plurality of transmission channels.

9. The method of claim 7, wherein the CSI includes information related to transmission characteristics from transmit antennas to the receive antennas.

10. The method of claim 7, wherein the CSI includes eigenmode information related to transmission characteristics from transmit antennas to the receive antennas.

11. The method of claim 6, further comprising:

grouping transmission channels having similar transmission capabilities to segments, and

wherein the puncturing is performed for each segment.

12. The method of claim 11, further comprising:

assigning a group of coded bits to each segment, and

wherein the puncturing is performed on the group of coded bits assigned to each segment.

13. The method of claim 11, wherein each segment includes transmission channels having SNR within a particular SNR range.

14. The method of claim 1, wherein the encoding is achieved via a Turbo code.

15. The method of claim 14, wherein the encoding provides a plurality of tail and parity bits for the plurality of information bits, and wherein the puncturing is performed on the plurality of tail and parity bits.

16. The method of claim 14, wherein the puncturing is performed such that unpunctured tail and parity bits are approximately evenly distributed over the plurality of information bits.

17. The method of claim 14, wherein the Turbo code includes two constituent codes operative to provide two streams of tail and parity bits, and wherein the puncturing is performed such that approximately equal number of tail and parity bits are deleted from the two streams of tail and parity bits.

18. The method of claim 1, wherein the coding rate for each transmission channel is selected to be between, and inclusive of, $n/(n+1)$ and $n/(n+2)$, where n is the number of information bits per modulation symbol supported by the transmission channel.

19. The method of claim 1, wherein the coding rate for each transmission channel is 1/2 or higher.

20. The method of claim 1, wherein the encoding is achieved via a convolutional code.

21. The method of claim 1, wherein the encoding is achieved via a block code.

22. The method of claim 1, further comprising:

inserting padding bits to fill available but unfilled bit positions in the plurality of transmission channels.

23. The method of claim 1, further comprising:

repeating at least some of the coded bits to fill available but unfilled bit positions in the plurality of transmission channels.

24. The method of claim 1, further comprising:

interleaving the plurality of coded bits.

25. The method of claim 24, wherein the puncturing is performed on interleaved coded bits.

26. The method of claim 24, wherein the encoding is achieved via a Turbo code comprised of two constituent codes, and wherein the plurality of information bits, a plurality of tail and parity bits from a first constituent code, and a plurality of tail and parity bits from a second constituent code are separately interleaved.

27. The method of claim 1, further comprising:

forming non-binary symbols for the plurality of transmission channels, wherein each non-binary symbol includes a group of unpunctured coded bits; and

mapping each non-binary symbol to a respective modulation symbol.

28. The method of claim 27, further comprising:

interleaving the plurality of coded bits, and

wherein the non-binary symbols are formed from the interleaved coded bits.

29. The method of claim 27, wherein the modulation scheme for each transmission channel is associated with a respective signal constellation having a plurality of points, and wherein each modulation symbol is representative of a particular point in the signal constellation for the modulation scheme.

30. The method of claim 29, wherein the plurality of points in each signal constellation are assigned with values based on a particular Gray mapping scheme.

31. The method of claim 30, wherein the values are assigned to the plurality of points in each signal constellation such that values for adjacent points in the signal constellation differ by one bit position.

32. The method of claim 1, further comprising:

adapting to changes in the plurality of transmission channels by repeating the determining the number of information bits per modulation symbol, the identifying the modulation scheme, and the determining the coding rate.

33. The method of claim 1, wherein the modulation scheme for each transmission channel supports transmission of two or more coded bits per modulation symbol.

34. The method of claim 1, wherein the transmission on the plurality of transmission channels are intended for a single recipient receiving device.

35. In an orthogonal frequency division modulation (OFDM) communication system, a method for preparing data for transmission on a plurality of transmission channels, wherein each transmission channel is operative to transmit a respective sequence of modulation symbols, the method comprising:

determining a number of information bits per modulation symbol supported by each transmission channel;

identifying a modulation scheme for each transmission channel such that the determined number of information bits per modulation symbol is supported;

determining a coding rate for each transmission channel based at least on the determined number of information bits per modulation symbol and the identified modulation scheme for the transmission channel, wherein at least two transmission channels are associated with different coding rates;

encoding a plurality of information bits in accordance with a particular Turbo code to provide a plurality of tail and parity bits;

interleaving the plurality of information and tail and parity bits in accordance with a particular interleaving scheme;

puncturing the plurality of interleaved bits in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the plurality of transmission channels, wherein the puncturing is adjusted to achieve the different coding rates for the at least two transmission channels;

forming non-binary symbols for the plurality of transmission channels, wherein each non-binary symbol includes a group of unpunctured coded bits; and

mapping each non-binary symbol to a respective modulation symbol.

36. A wireless communication system operative to transmit data on a plurality of transmission channels, wherein each transmission channel is used to transmit a respective sequence of modulation symbols, the system comprising:

an encoder configured to encode a plurality of information bits in accordance with a particular encoding scheme to provide a plurality of coded bits, and to puncture the plurality of coded bits in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the plurality of transmission channels, wherein each transmission channel is capable of transmitting a particular number of information bits per modulation symbol via a particular modulation scheme selected for the transmission channel, wherein each transmission channel is further associated with a particular coding rate based at least on the number of information bits per modulation symbol supported by the transmission channel and its modulation scheme, wherein at least two transmission channels are associated with different coding rates, and wherein the encoder is further configured to adjust the puncturing to achieve the different coding rates for the at least two transmission channels.

37. The system of claim 36, further comprising:

a channel interleaver coupled to the encoder and configured to interleave the plurality of coded bits, and

wherein the encoder is configured to puncture the interleaved bits.

38. The system of claim 37, further comprising:

a symbol mapping element coupled to the channel interleaver and configured to form non-binary symbols for the plurality of transmission channels, and to map each non-binary symbol to a respective modulation symbol, wherein each non-binary symbol includes a group of unpunctured coded bits.

39. The system of claim 38, further comprising:

a signal processor coupled to the symbol mapping element and configured to pre-condition the modulation symbols for the plurality of transmission channels to implement a multiple-input multiple-output (MIMO) transmission.

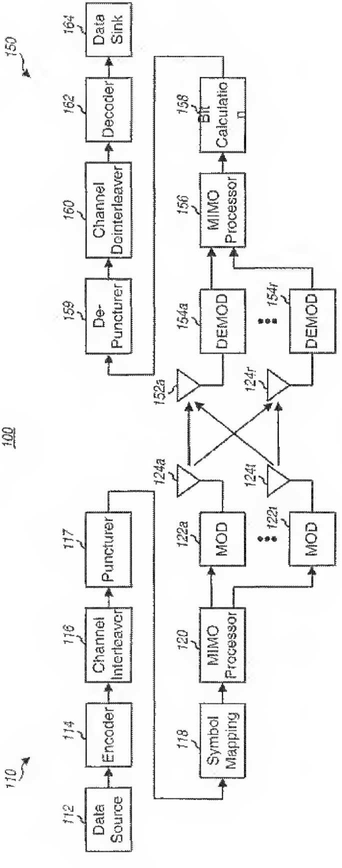
100
110
112
Data Source
114
Encoder
116
Channel Interleaver
117
Puncturer
118
Symbol Mapping
120
MIMO Processor
122a
MOD
122t
MOD
124a
124t
152a
124r
150
154a
DEMOD
154r
DEMOD
156
MIMO Processor
158
Bit Calculatio n
159
De-Puncturer
160
Channel Deinterleaver
162
Decoder
164
Data Sink

FIG. 1

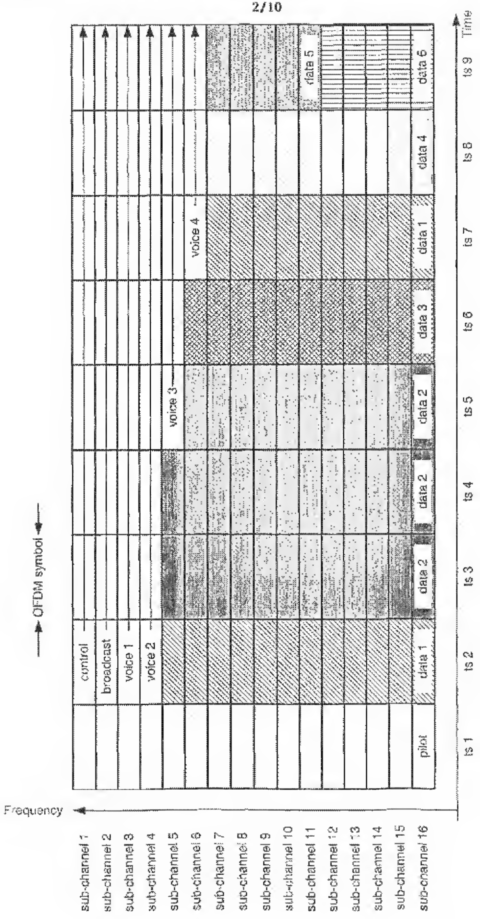
OFDM symbol
control
broadcast
voice 1
voice 2
voice 3
voice 4
pilot
data 1
data 2
data 2
data 2
data 3
data 1
data 4
data 5
data 6
ts 1
ts 2
ts 3
ts 4
ts 5
ts 6
ts 7
ts 8
ts 9
Time
Frequency
sub-channel 1
sub-channel 2
sub-channel 3
sub-channel 4
sub-channel 5
sub-channel 6
sub-channel 7
sub-channel 8
sub-channel 9
sub-channel 10
sub-channel 11
sub-channel 12
sub-channel 13
sub-channel 14
sub-channel 15
sub-channel 16

FIG. 2

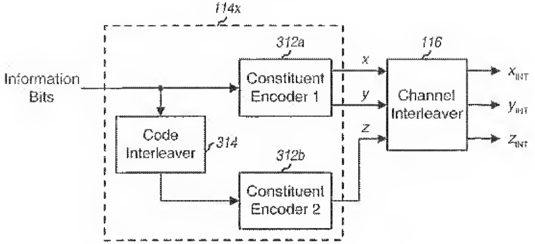
114x
312a
116
Information Bits
Constituent Encoder 1
x
y
z
Channel Interleaver
x_{INT}
y_{INT}
z_{INT}
Code Interleaver
314
312b
Constituent Encoder 2

FIG. 3A

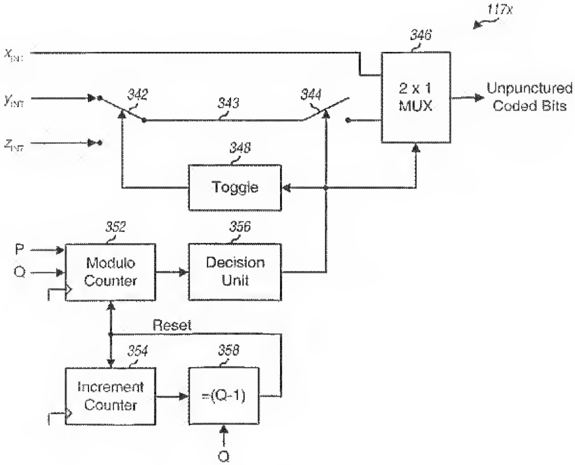
117x
346
x_{INT}
y_{INT}
z_{INT}
342
343
344
2 x 1 MUX
Unpunctured Coded Bits
348
Toggle
352
356
P
Q
Modulo Counter
Decision Unit
Reset
354
358
Increment Counter
=(Q-1)
Q

FIG. 3C

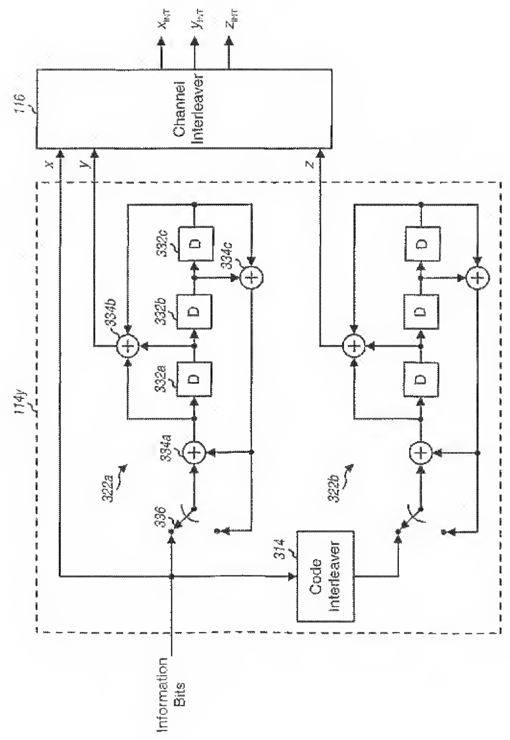
116
Channel Interleaver
X_{INT}
Y_{INT}
Z_{INT}
x
y
z
114y
332a
332b
332c
334a
334b
334c
D
D
D
D
D
D
322a
322b
336
314
Code Interleaver
Information Bits

FIG. 3B

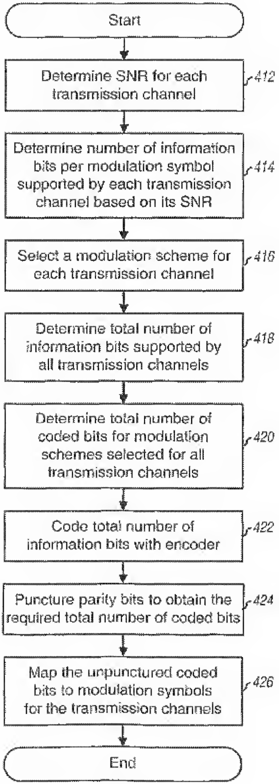
Start
Determine SNR for each transmission channel
412
Determine number of information bits per modulation symbol supported by each transmission channel based on its SNR
414
Select a modulation scheme for each transmission channel
416
Determine total number of information bits supported by all transmission channels
418
Determine total number of coded bits for modulation schemes selected for all transmission channels
420
Code total number of information bits with encoder
422
Puncture parity bits to obtain the required total number of coded bits
424
Map the unpunctured coded bits to modulation symbols for the transmission channels
426
End

FIG. 4A

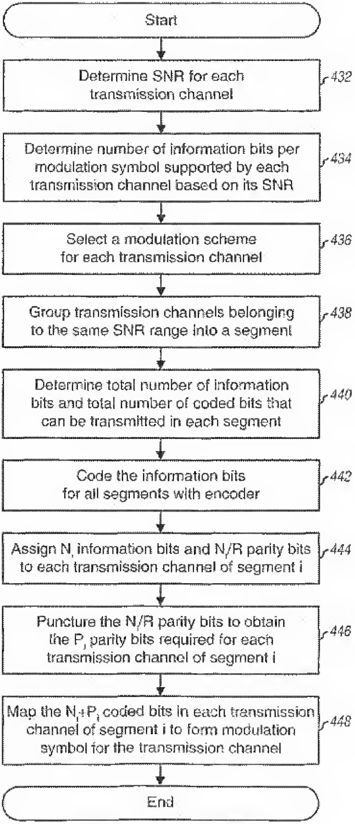
Start
Determine SNR for each transmission channel
432
Determine number of information bits per modulation symbol supported by each transmission channel based on its SNR
434
Select a modulation scheme for each transmission channel
436
Group transmission channels belonging to the same SNR range into a segment
438
Determine total number of information bits and total number of coded bits that can be transmitted in each segment
440
Code the information bits for all segments with encoder
442
Assign N_i information bits and N_i/R parity bits to each transmission channel of segment i
444
Puncture the N_i/R parity bits to obtain the P_i parity bits required for each transmission channel of segment i
446
Map the N_i+P_i coded bits in each transmission channel of segment i to form modulation symbol for the transmission channel
448
End

FIG. 4B

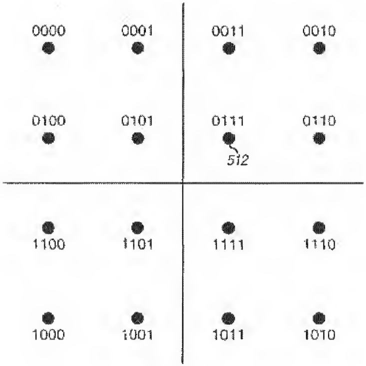
0000
0001
0011
0010
0100
0101
0111
0110
512
1100
1101
1111
1110
1000
1001
1011
1010

FIG. 5

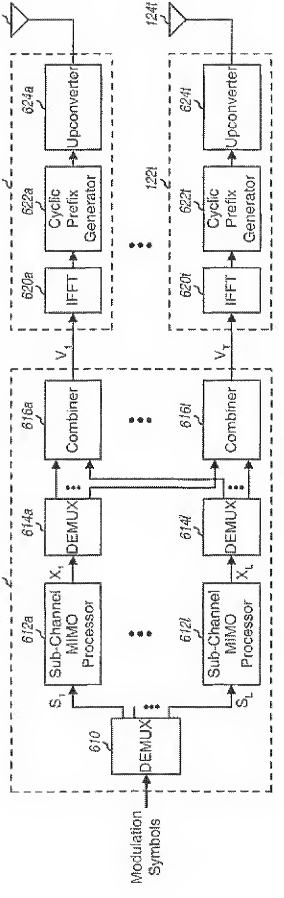
124a
124t
122a
122t
624a
Upconverter
624t
Upconverter
622a
Cyclic Prefix Generator
622t
Cyclic Prefix Generator
620a
IFFT
620t
IFFT
V_1
V_T
616a
Combiner
616t
Combiner
614a
DEMUX
614t
DEMUX
120x
X_1
X_L
612a
Sub-Channel MIMO Processor
612t
Sub-Channel MIMO Processor
S_1
S_L
610
DEMUX
Modulation Symbols

FIG. 6

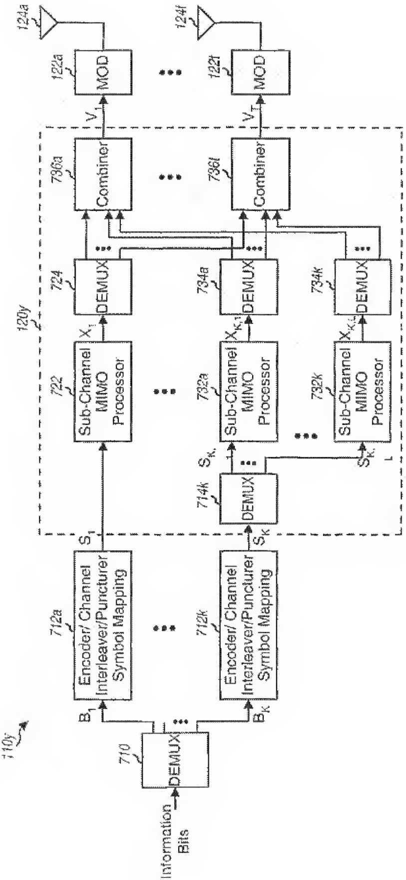
110y
Information Bits
710
DEMUX
B_1
B_K
712a
Encoder/ Channel Interleaver/Puncturer Symbol Mapping
712k
Encoder/ Channel Interleaver/Puncturer Symbol Mapping
S_1
S_K
120y
714k
DEMUX
$S_{K,1}$
$S_{K,L}$
722
Sub-Channel MIMO Processor
732a
Sub-Channel MIMO Processor
732k
Sub-Channel MIMO Processor
X_1
$X_{K,1}$
$X_{K,L}$
724
DEMUX
734a
DEMUX
734k
DEMUX
736a
Combiner
736t
Combiner
V_1
V_T
122a
MOD
122t
MOD
124a
124t

FIG. 7

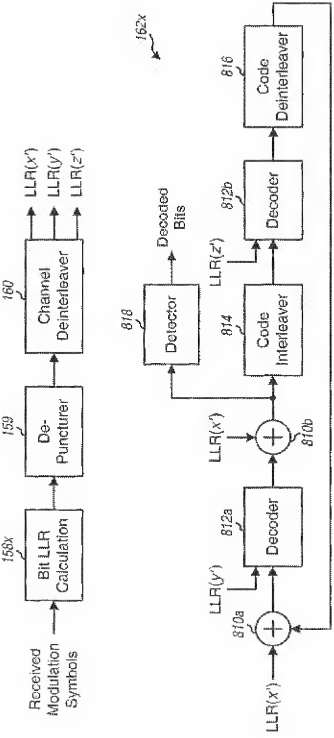
162x
Received Modulation Symbols
158x
Bit LLR Calculation
159
De-Puncturer
160
Channel Deinterleaver
LLR(x')
LLR(y')
LLR(z')
LLR(x')
810a
LLR(y')
812a
Decoder
LLR(x')
810b
818
Detector
Decoded Bits
814
Code Interleaver
LLR(z')
812b
Decoder
816
Code Deinterleaver

FIG. 8

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/02143

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/00 H04L27/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, COMPENDEX, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X A	US 5 197 061 A (LE FLOCH BERNARD ET AL) 23 March 1993 (1993-03-23) column 2, line 67 -column 3, line 10 column 3, line 15 - line 18 column 4, line 29 - line 40 column 4, line 59 - line 62 column 5, line 18 - line 21 column 5, line 54 - line 68 column 6, line 40 - line 47 column 9, line 51 - line 59 --- -/--	1,3,6, 35,36 2,4,5, 7-34, 37-39

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
3 July 2002	23/07/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Schiffer, A

Form PCT/ISA/210 (second sheet) (July 1992)

International Application No
PCT/US 02/02143

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	MATSUOKA H ET AL: "Adaptive modulation system with variable coding rate concatenated code for high quality multi-media communication systems" VEHICULAR TECHNOLOGY CONFERENCE, 1996. MOBILE TECHNOLOGY FOR THE HUMAN RACE., IEEE 46TH ATLANTA, GA, USA 28 APRIL-1 MAY 1996, NEW YORK, NY, USA,IEEE, US, 28 April 1996 (1996-04-28), pages 487-491, XP010162435 ISBN: 0-7803-3157-5	1,6-8, 30,35,36
A		2-5, 9-29, 31-34, 37-39
	abstract page 487, right-hand column, paragraph 1 page 487, right-hand column, paragraph 3 page 488, left-hand column, paragraph 1 page 488, left-hand column, paragraph 3 - paragraph 4 page 488, right-hand column, paragraph 5	
A	WO 00 27085 A (BROADCOM CORP ;JAFFE STEVEN (US); LIN THUJI SIMON (US); JOSHI ROBI) 11 May 2000 (2000-05-11) page 2, line 34 -page 3, line 2 page 3, line 11 - line 15 page 3, line 36 -page 4, line 20 page 4, line 2 - line 4 page 4, line 22 - line 29 page 5, line 31 - line 37 page 8, line 32 - line 37	1-39
A	MUNETA S ET AL: "A NEW FREQUENCY-DOMAIN LINK ADAPTATION SCHEME FOR BROADBAND OFDM SYSTEMS" VTC 1999-FALL. IEEE VTS 50TH. VEHICULAR TECHNOLOGY CONFERENCE. GATEWAY TO THE 21ST. CENTURY COMMUNICATIONS VILLAGE. AMSTERDAM, SEPT. 19 - 22, 1999, IEEE VEHICULAR TECHNOLGY CONFERENCE, NEW YORK, NY: IEEE, US, vol. 1 CONF. 50, September 1999 (1999-09), pages 253-257, XP000929050 ISBN: 0-7803-5436-2 abstract page 253, right-hand column, paragraph 2 - paragraph 3 page 254, left-hand column, paragraph 1 page 254, left-hand column, paragraph 3 page 254, right-hand column, paragraph 2 page 255, left-hand column, paragraph 3	1-39

-/--

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	SAMPEI S ET AL: "ADAPTIVE MODULATION/TMDA SCHEME FOR LARGE CAPACITY PERSONAL MULTI-MEDIA COMMUNICATION SYSTEMS" IEICE TRANSACTIONS ON COMMUNICATIONS, INSTITUTE OF ELECTRONICS INFORMATION AND COMM. ENG. TOKYO, JP, vol. E77-B, no. 9, 1 September 1994 (1994-09-01), pages 1096-1103, XP000474107 ISSN: 0916-8516 page 1096, right-hand column, paragraph 3 page 1097, left-hand column, paragraph 3 page 1097, right-hand column, paragraph 5 page 1098, right-hand column, paragraph 6 page 1100, left-hand column, paragraph 2	1-39

INTERNATIONAL SEARCH REPORT

Information on patent family members

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Patent document cited in search report		Publication date	Patent family member(s)		Publication date
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(54) Title: METHOD AND APPARATUS FOR A RATE CONTROL IN A HIGH DATA RATE COMMUNICATION SYSTEM

(57) Abstract: A method and an apparatus for rate control in a high data rate (HDR) communication system are disclosed. An exemplary HDR communication system defines a set of data rates, at which an access point (AP) may send data packets to an access terminal (AT). The data rate is selected to maintain targeted packet error rate (PER). The AT's open loop algorithm measures received signal to interference and noise ratio (SINR) at regular intervals, and uses the information to predict an average SINR over the next packet duration. The AT's closed loop algorithm measures a packet error rate (PER) of the received signal, and uses the PER to calculate a closed loop correction factor. The loop correction factor is added to the SINR value predicted by the open loop, resulting in an adjusted SINR. The AT maintains a look up table, which comprises a set of SINR thresholds that represent a minimum SINR necessary to successfully decode a packet at each data rate. The AT uses the adjusted set of SINR thresholds in the look up table to select the highest data rate, the SINR threshold of which is below the predicted SINR. The AT then requests, over the reverse link, that the AP send the next packet at this data-rate.

METHOD AND APPARATUS FOR A RATE CONTROL IN A HIGH DATA RATE COMMUNICATION SYSTEM

BACKGROUND OF THE INVENTION

I. Field of the Invention

The current invention relates to communication. More particularly, the present invention relates to a novel method and apparatus for adaptive rate selection in a wireless communication system.

II. Description of the Related Art

A modern communications system is required to support a variety of applications. One such communications system is a code division multiple access (CDMA) system that conforms to the "TIA/EIA/IS-95 Mobile Station-Base Station Compatibility Standard for Dual-Mode Wide-Band Spread Spectrum Cellular System," hereinafter referred to as the IS-95 standard. The CDMA system supports voice and data communication between users over a terrestrial link. The use of CDMA techniques in a multiple access communication system is disclosed in U.S. Patent No. 4,901,307, entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS," and U.S. Patent No. 5,103,459, entitled "SYSTEM AND METHOD FOR GENERATING WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM," both assigned to the assignee of the present invention and incorporated herein by reference.

In a CDMA system, communications between users are conducted through one or more base stations. In wireless communication systems, forward link refers to the channel through which signals travel from a base station to a subscriber station, and reverse link refers to channel through which signals travel from a subscriber station to a base station. By transmitting data on a reverse link to a base station, a first user on one subscriber station may communicate with a second user on a second subscriber station. The base

station receives the data from the first subscriber station and routes the data to a base station serving the second subscriber station. Depending on the location of the subscriber stations, both may be served by a single base station or multiple base stations. In any case, the base station serving the second subscriber station sends the data on the forward link. Instead of communicating with a second subscriber station, a subscriber station may also communicate with a wireline telephone through a public switched telephone network (PSTN) coupled to the base station, or a terrestrial Internet through a connection with a serving base station.

Given the growing demand for wireless data applications, the need for very efficient wireless data communication systems has become increasingly significant. The IS-95 standard specifies transmitting traffic data and voice data over the forward and reverse links. A method for transmitting traffic data in code channel frames of fixed size is described in detail in U.S. Patent No. 5,504,773, entitled "METHOD AND APPARATUS FOR THE FORMATTING OF DATA FOR TRANSMISSION", assigned to the assignee of the present invention and incorporated by reference herein. In accordance with the IS-95 standard, the traffic data or voice data is partitioned into code channel frames that are 20 milliseconds wide with data rates as high as 14.4 Kbps.

In mobile radio communication systems, there are significant differences between the requirements for providing voice and data services (i.e., non-voice services such as Internet or fax transmissions). Unlike data services, voice services require stringent and fixed delays between speech frames. Typically, the overall one-way delay of speech frames used for transmitting voice information must be less than 100 msec. By contrast, transmission delays that occur during data (i.e., non-voice information) services can vary and larger delays then those that can be tolerated for voice services can be utilized.

Another significant difference between voice and data services is that, in contrast to data services, voice services require a fixed and common grade of service. Typically, for digital systems providing voice services, this requirement is met by using a fixed and equal transmission rate for all users

and a maximum tolerable error rate for speech frames. For data services, the grade of service can vary from user to user.

Yet another difference between voice services and data services is that voice services require a reliable communication link which, in the case of a CDMA communication system, is provided using a soft handoff. A soft handoff requires the redundant transmission of the same voice information from two or more base stations to improve reliability. A soft handoff method is disclosed in U.S. Patent No. 5,101,501, entitled "SOFT HANDOFF IN A CDMA CELLULAR TELEPHONE SYSTEM." This additional reliability is not required to support data services, because data packets received in error can be retransmitted.

As a mobile station moves in a mobile radio communication system, the quality of the forward link (and the capacity of the forward link to transmit data) will vary. Thus, at some moments a given forward link between a base station and a mobile station will be able to support a very high data transmission and, at other moments, the same forward link may only be able to support a much reduced data transmission rate. In order to maximize the throughput of information on the forward link, it would be desirable if the transmission of data on the forward link could be varied so as to increase the data rate during those intervals where the forward link can support a higher transmission rate.

When non-voice traffic is being sent from a base station to a mobile station on a forward link, it may be necessary to send control information from the mobile station to the base station. At times, however, even though the forward link signal may be strong, the reverse link signal may be weak, thereby resulting in a situation where the base station cannot receive control information from the mobile station. In such situations, where the forward link and the reverse link are unbalanced, it may be undesirable to increase the transmit power on the reverse link in order to improve the reception quality of the control information at the base station. For example, in CDMA systems, increasing the transmit power on the reverse link would be undesirable, as such

a power increase could adversely affect the reverse link capacity seen by other mobile stations in the system. It would be desirable to have a data transmission system where the forward and reverse links associated with each mobile station were maintained in a balanced state without adversely impacting the reverse link capacity. It would be further desirable if such a system could maximize the throughput of non-voice data on individual forward links when such links are sufficiently strong to support higher data rates.

One approach to the aforementioned requirements in high data rate (HDR) systems is to keep the transmit power fixed and vary the data rate depending on the users' channel conditions. Consequently, in a modern HDR system, Access Point(s) (APs) always transmit at maximum power to only one Access Terminal (AT) in each time slot, and the AP uses rate control to adjust the maximum rate that the AT can reliably receive. An AP is a terminal allowing high data rate transmission to ATs.

As used in this document, a time slot is a time interval of finite length, e.g., 1.66 ms. A time slot can contain one or more packets. A packet is a structure, comprising a preamble, a payload, and a quality metric, e.g., a cyclical redundancy check (CRC). The preamble is used by an AT to determine whether a packet has been intended for the AT.

An exemplary HDR system defines a set of data rates, ranging from 38.4kbps to 2.4 Mbps, at which an AP may send data packets to an AT. The data rate is selected to maintain a targeted packet error rate (PER). The AT measures the received signal to interference and noise ratio (SINR) at regular intervals, and uses the information to predict an average SINR over the next packet duration. An exemplary prediction method is disclosed in co-pending application serial number 09/394,980 entitled "SYSTEM AND METHOD FOR ACCURATELY PREDICTING SIGNAL TO INTERFERENCE AND NOISE RATIO TO IMPROVE COMMUNICATIONS SYSTEM PERFORMANCE," assigned to the assignee of the present invention and incorporated herein by reference.

FIG. 1 shows a conventional open loop rate control apparatus **100**. A stream of past SINR values at instances [n-m], . . .[n-1], [n], each measured over a duration of a corresponding packet, is provided to a predictor **102**. The predictor **102** predicts the average SINR over the next packet duration in accordance with the following equation:

$$OL\_SINR_{Predicted} = OL\_SINR_{Estimated} - K \cdot \sigma_e \qquad (1)$$

In Equation (1), $OL\_SINR_{Predicted}$ is a SINR predicted by the open loop for the next packet, $OL\_SINR_{Estimated}$ is a SINR estimated by the open loop based on past SINR values, K is a back-off factor, and σ_e is a standard deviation of an error metric.

The estimated SINR may be obtained, for example, by selecting an output from a bank of low pass filters acting on past measurements of SINR. Selection of a particular filter from the filter bank may be based on an error metric, defined as a difference between the particular filter output and measured SINR over a packet duration immediately following the output. The predicted SINR is obtained by backing off from the filter output by an amount equal to the product of the back-off factor K and the standard deviation σ_e of the error metric. The value of the back-off factor K is determined by a back-off control loop, which ensures that a tail probability, i.e., probability that predicted SINR exceeds the measured SINR, is achieved for a certain percentage of time.

The $SINR_{Predicted}$ value is provided to a look up table **104** that maintains a set of SINR thresholds that represent the minimum SINR required to successfully decode a packet at each data rate. An AT (not shown) uses the look up table **104** to select the highest data rate whose SINR threshold is below the predicted SINR, and requests that an AP (not shown) send the next packet at this datarate.

The aforementioned method is an example of an open loop rate control method that determines the best rate at which to receive the next packet, based

only on the measurement of the channel SINR, without any information about the decoder error rate (for packets of each data rate) at a given SINR under the prevailing channel conditions. Any open loop rate control algorithm suffers from several shortcomings, some of which are discussed below. First, a certain tail probability, e.g., 2%, does not imply a PER of 2%. This is because PER is a monotonically decreasing function of SINR, with a finite slope that depends on the coding scheme and channel conditions. However, Equation (1) assumes "brick wall" PER characteristics, i.e., a packet is guaranteed to be decoded whenever the SINR exceeds the threshold for the corresponding rate, and a packet is in error whenever the SINR falls below the threshold. Furthermore, the open loop rate control method uses a fixed set of SINR thresholds, which ensures packet error rates close to the target error rate under worst-case channel conditions. However, the performance of the decoder depends not only on the SINR, but also on channel conditions. In other words, a method that uses a fixed set of SINR thresholds for all channels achieves different packet error rates on different channels. Consequently, while the open loop method works optimally under the worst-case channel conditions, it is possible that under typical channel conditions, the method results in much lower error rates than is necessary, at the expense of diminished throughput. Additionally, a practical rate control method necessitates a small, finite set of data rates. The rate selection method always selects the nearest lower data rate in order to guarantee an acceptable PER. Thus, rate quantization results in loss of system throughput.

Therefore, there exists a need to address deficiencies of the existing method.

SUMMARY OF THE INVENTION

The present invention is directed to a novel method and apparatus for adaptive rate selection in a wireless communication system. Accordingly, in one aspect of the invention, SINR predicted by an open loop method is modified by a closed loop correction. The closed loop correction is updated in accordance with packet error events and a target error rate.

In another aspect of the invention, the closed loop correction is advantageously updated in accordance with a frequency with which packets are received.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 illustrates a block diagram of a conventional, open loop rate-control apparatus.

FIG. 2 illustrates a block diagram of an apparatus for a rate control method in accordance with one embodiment of the invention.

FIG. 3 illustrates a flowchart of an exemplary method of updating an outer loop correction.

DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

FIG. 2 illustrates an exemplary communication system **200** capable of implementing embodiments of the invention. An AP **204** transmits signals to an AT **202** over a forward link **206a**, and receives signals from the AT **202** over a reverse link **206b**. The communication system **200** can be operated bi-directionally, each of the terminals **202, 204** operating as a transmitter unit or a receiver unit, or both concurrently, depending on whether data is being transmitted from, or received at, the respective terminal **202, 204**. In a cellular wireless communication system embodiment, the transmitting terminal **204** can be a base station (BS), the receiving terminal **202** can be a mobile station (MS), and the forward link **206a** and reverse link **206b** can be electromagnetic spectra.

The AT **202** contains an apparatus for a rate control method in accordance with one embodiment of the present invention. The apparatus contains two control loops, an open loop and a closed loop.

The open loop, comprising a SINR predictor **208** and a look up table **210**, controls the forward-link data rate based on the difference between the average SINR of the next packet and SINR thresholds of all the data rates. A signal arriving at the AT **202** from the AP **204** over the forward link **206a** in packets is provided to a decoder **212**. The decoder **212** measures an average SINR over the duration of each packet, and provides the SINRs to the SINR predictor **208**.

In one embodiment, the predictor **208** predicts a SINR ($OL\_SNIR_{Predicted}$) value of the next packet in accordance with Equation (1). However, one skilled in the art will understand that any open loop method, not limited to the one expressed by Equation (1), may be used. The $OL\_SNIR_{Predicted}$ value is provided to the look up table **210**. The look up table **210** maintains a set of SINR thresholds that represents the minimum SINR required to successfully decode a packet at each data rate. The set of SINR thresholds is adjusted by the operation of the closed loop.

The closed loop utilizes PER information provided by the decoder **212** to determine a closed loop correction value L in block **214**. The closed loop correction value L adjusts the set of SINR thresholds in the look up table **204** in accordance with the following equation:

$$CL\_SINR_{Predicted} = OL\_SINR_{Predicted} + L, \tag{2}$$

In equation (2), L represents the closed loop correction to the open loop prediction of SINR over the next packet duration. Adding L to the SINR predicted by the open loop algorithm in Equation (1) is equivalent to subtracting L from the SINR thresholds used for rate control. Because the correction term L is updated in accordance with PER information, which reflects the prevailing channel conditions, the set of SINR thresholds is better matched to the prevailing channel conditions.

The AT **202** uses the adjusted set of SINR thresholds in the look up table **210** to select the highest data rate, the SINR threshold of which is below the predicted SINR. The AT **202** then requests, over the reverse link **206b**, that the AP **204** sends the next packet at this data-rate.

Although the predictor **208**, the decoder **212**, and the closed loop correction block **214** are shown as separate elements, one skilled in the art will appreciate that the physical distinction is made for explanatory purposes only. The predictor **208**, the decoder **212**, and the closed loop correction block **214** may be incorporated into a single processor accomplishing the above-mentioned processing. Thus, the processor may be, e.g., a general-purpose processor, a digital signal processor, a programmable logic array, and the like. Furthermore, the look up table **210** is a space in a memory. The memory may be a part of the above-mentioned processor or processors, or be a separate element. The implementation of the memory is a design choice. Thus, the memory can be any media capable of storing information, e.g., a magnetic disk, a semiconductor integrated circuit, and the like.

FIG. 3 illustrates a flowchart of an exemplary method of updating L to ensure the best possible throughput with acceptable error rates.

In step **300**, a normalized activity factor (AF) variable is initialized by an AT (not shown) to a value of zero or one. The AF quantifies a time fraction for which the AT receives packets on the forward link. An AF being equal to one implies that the AT **202** is receiving packets most of the time, whereas an AF being equal to zero implies that the forward link to the given AT is mostly idle. In one embodiment, the AF is initialized at the instant when the AT initiates a new communication. In that case, it may be advantageous to initialize the AF to one because the AT is receiving packets. The AF is updated at the end of each time slot according to the following equations:

$$AF_{New} = (1 - f) \cdot AF_{Old} + f\,, \tag{3}$$

or

$$AF_{New} = (1 - f) \cdot AF_{Old}\,, \tag{4}$$

where:

$f \in (0,1)$ is a parameter controlling a rate of change of the AF. In one embodiment of the invention, f is set to 1/50.

Equation (3) is used when the AT finds a packet preamble at the beginning of a time slot, or is still demodulating a packet whose preamble was detected in an earlier time slot. This happens when the AT sends a request for data, and an AP (not shown) sends the requested data. Equation (4) is used when the AT is not in the middle of packet demodulation, searches for a packet preamble, and fails to find the preamble. This happens when the AT sends a request for data, and the AP fails to receive or ignores the request for data, and decides to serve some other AT in the system.

In step **300**, the outer loop correction variable L is also initialized by the AT. L can be initialized to any value between L_{min} and L_{max}. L_{min}, L_{max} may attain any value. Exemplary values are cited below. In one embodiment, L is initialized to 0 dB.

In step **300**, a mode of operation is also initialized. There are two modes: a normal mode and a fast attack mode. The motivation behind defining the two modes for the rate control algorithm is based on the knowledge that an optimal step size for upward and downward corrections of L depends on a target PER, the packet arrival process, and preamble false alarm statistics. While the

preamble false alarm statistics are relatively constant and correlated with the outer loop term L, the packet arrival process is time varying and unknown *apriori* at the AT. As discussed above, data traffic tends to be bursty, with an idle state characterized by infrequent packet arrival, and busy, with frequent packet arrival. Consequently, the normal mode is used during steady state. Fast attack mode designed to recover quickly from long periods of inactivity is used when preamble false alarms tend to drive the rate control algorithm toward the conservative regime.

The rules for determining the mode of the algorithm, as well as the rules for updating L, are based on the detection of good or bad packets. The access terminal is said to receive a good packet if it detects the packet preamble, demodulates and decodes the packet, and recovers a valid CRC. The access terminal is said to receive a bad packet if it detects a packet preamble, but upon demodulating and decoding the packet, it obtains an invalid CRC.

The transition to the fast attack mode occurs if all the following conditions are satisfied:

$$L < L_{AMThreshold}\,, \tag{5}$$

$$AF < AF_{Idle}\,, \text{ and} \tag{6}$$

the two most recently received packets are good.

In Equations (5)-(6), $L_{AMThreshold}$ is a threshold controlling the transition to the fast attack mode with respect to L. In one embodiment of the invention, the $L_{AMThreshold}$ threshold is set to 0dB. AF_{Idle} is a threshold controlling the transition to the fast attack mode with respect to AF. In one embodiment of the invention, the AF_{Idle} threshold is set to 10%.

The transition to the normal mode occurs if any of the following conditions are satisfied:

$$L \geq L_{NMThreshold}\,, \tag{7}$$

$$AF \geq AF_{Busy}\,, \text{ or} \tag{8}$$

the most recently received packet is bad.

In Equations (7)-(8), $L_{NMThreshold}$ is a threshold controlling the transition to the normal mode with respect to L. In one embodiment of the invention, the $L_{NMThreshold}$ threshold is set to 2dB. AF_{Busy} is a threshold controlling the transition to the normal mode with respect to A. In one embodiment of the invention, the AF_{Busy} threshold is set to 25%.

Upon finishing initialization, the AT waits for a new time slot. Once a time slot is detected in step **302**, the AF is updated in step **304** using Equations (3) or (4), and the mode is updated in step **306** using Equations (5)-(6) or (7)-(8).

In step **308**, a test is made whether the slot belonged to a new packet. If a new packet has not been detected, the method returns to step **302**. If a new packet has been detected, the packet is tested in step **310**, and if a bad packet has been detected, the method continues in step **312**. In step **312**, the value of L is updated in accordance with the following equation:

$$L_{new} = \max(L_{old} - \delta, L_{\min}), \qquad (9)$$

where δ is a step size. In one embodiment of the invention, the step size is set to 0.25 dB. L_{min} is the minimum value that L can attain. In one embodiment of the invention, the value of L_{min} is limited to -1 dB. The method then returns to step **302**.

If, in step **310**, a good packet was detected, the method continues in step **314**. In step **314**, the mode is tested. If the AT is in fast attack mode, the value of **L** is updated in accordance with the following equation in step **316**:

$$L_{new} = \min(L_{Old} + \delta', L_{\max}), \qquad (9)$$

where:

δ' is a step size. In one embodiment of the invention, the step size is set to 0.25 dB. L_{max} is the maximum value that L can attain. In one embodiment of the invention, the value of L_{max} is limited to 3 dB. Once L is updated in step **318** the method returns to step **302**.

If a normal mode was detected in step **314**, the method continues in step **318**, where the value of L is updated in accordance with the following equation:

$$L_{new} = \min(L_{old} + TARGET\_PER \cdot \delta, L_{\max}). \qquad (10)$$

In Equation (9), δ is a step size. In one embodiment of the invention, the step size is set to 0.25 dB. *TARGET_PER* is the PER to be maintained. L_{max} is the maximum value that **L** can attain. In one embodiment of the invention, the value of L_{max} is limited to 3 dB. Once L is updated in step **318** the method returns to step **302**.

The previous description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. The

various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. A method for rate selection in a wireless communication system comprising the steps of:

determining an open loop prediction of signal to noise and interference ratio;

determining a closed loop correction; and

selecting a data rate in accordance with said open loop prediction and said closed loop correction.

2. The method of claim 1 wherein said step of determining a closed loop correction comprises the steps of:

determining a quality of a received packet; and

decreasing said closed loop correction if said quality is bad.

3. The method of claim 2 wherein said step of decreasing is carried out in accordance with an equation:

$$L_{new} = \max(L_{old} - \delta, L_{\min}),$$

wherein L_{new} is an updated value of said outer loop correction, L_{old} is a previous value of said outer loop correction, δ is a step size, and L_{min} is the minimum value that said outer loop correction can attain.

4. The method of claim 1 wherein said step of determining a closed loop correction comprises the steps of:

determining a quality of a received packet; and

increasing said closed loop correction if said quality is good.

5. The method of claim 4 wherein said step of increasing comprises the steps of:

determining a mode of operation; and

increasing said closed loop correction in accordance with said mode of operation.

6. The method of claim 5 wherein said step of determining a mode of operation comprises the steps of:

determining a time fraction for which a packet is received; and

selecting said mode of operation in accordance with said time fraction.

7. The method of claim 6 wherein when a packet is detected said step of determining is carried out in accordance with an equation:

$$AF_{New} = (1 - f) \cdot AF_{Old} + f$$

wherein AF_{new} is an updated value of said time fraction, L_{old} is a previous value of said time fraction, and $f \in (0,1)$ is a parameter controlling a rate of change of said time fraction.

8. The method of claim 6 wherein when a packet detection fails said step of determining is carried out in accordance with an equation:

$$AF_{New} = (1 - f) \cdot AF_{Old},$$

wherein AF_{new} is an updated value of said time fraction, L_{old} is a previous value of said time fraction, and $f \in (0,1)$ is a parameter controlling a rate of change of said time fraction.

9. The method of claim 6 wherein the step of selecting comprises the step of selecting a fast attack mode if all of the following conditions are satisfied:

$L < L_{AMThreshold}$,
$AF < AF_{Idle}$,
the two most recently received packets are good,

wherein $L_{AMThreshold}$ is a threshold controlling the transition to said fast attack mode with respect to L and AF_{Idle} is a threshold controlling the transition to said fast attack mode with respect to AF.

10. The method of claim 6 wherein the step of selecting comprises the step of selecting a normal mode if any of the following conditions are satisfied:

$L \geq L_{NMThreshold}$,
$AF \geq AF_{Busy}$, or
the most recently received packet is bad,

wherein $L_{NMthreshold}$ is a threshold controlling the transition to said normal mode with respect to L. AF_{Idle} is a threshold controlling the transition to said normal mode with respect to AF; and

11. The method of claim 5 wherein when in a fast attack mode of operation said step of increasing is carried out in accordance with an equation:

$$L_{new} = \min(L_{old} + \delta', L_{\max}),$$

wherein L_{new} is an updated value of said outer loop correction, L_{old} is a previous value of said outer loop correction, δ' is a step size, and L_{max} is the maximum value that said outer loop correction can attain.

12. The method of claim 5 wherein when in a normal mode of operation said step of increasing is carried out in accordance with an equation:

$$L_{new} = \min(L_{old} + TARGET\_PER \cdot \delta, L_{\max}),$$

wherein L_{new} is an updated value of said outer loop correction, L_{old} is a previous value of said outer loop correction, $TARGET\_PER$ is a packet error rate to be attained, δ' is a step size, and L_{max} is the maximum value that said outer loop correction can attain.

13. The method of claim 1 wherein said step of selecting comprises the steps of:

summing said open loop prediction of signal to noise and interference ratio and said closed loop correction; and

determining said data rate as the highest data rate, a signal to noise ratio of which is below said summed signal to noise ratio.

14. An apparatus for selecting rate in a wireless communication system, comprising:

a processor; and

a storage medium coupled to the processor and containing a set of instructions executable by the processor to:

determine an open loop prediction of signal to noise and interference ratio;

determine a closed loop correction; and

select a data rate in accordance with said open loop prediction and said closed loop correction.

15. The apparatus of claim 14 wherein said processor comprises a signal to noise and interference ratio predictor and a closed loop correction calculator.

16. The apparatus of claim 14 wherein said processor is configured to decrease said closed loop correction if a quality of a received packet is bad.

17. The apparatus of claim 16 wherein said processor is configured to decrease said closed loop correction in accordance with an equation:

$$L_{new} = \max(L_{old} - \delta, L_{\min}),$$

wherein L_{new} is an updated value of said outer loop correction, L_{old} is a previous value of said outer loop correction, δ is a step size, and L_{min} is the minimum value that said outer loop correction can attain.

18. The apparatus of claim 14 wherein said processor is configured to increase said closed loop correction if a quality of a received packet is good.

19. The apparatus of claim 18 wherein said processor is configured to:

determine a mode of operation; and

increase said closed loop correction in accordance with said mode of operation.

20. The apparatus of claim 19 wherein said processor is configured to:

determine a time fraction for which a packet is received; and

select said mode of operation in accordance with said time fraction.

21. The apparatus of claim 20 wherein when a packet is detected said processor is configured to determine said time fraction in accordance with an equation:

$$AF_{New} = (1 - f) \cdot AF_{Old} + f$$

wherein AF_{new} is an updated value of said time fraction, L_{old} is a previous value of said time fraction, and $f \in (0,1)$ is a parameter controlling a rate of change of said time fraction.

22. The apparatus of claim 20 wherein when a packet detection fails said processor is configured to determine said time fraction in accordance with an equation:

$$AF_{New} = (1-f) \cdot AF_{Old},$$

wherein AF_{new} is an updated value of said time fraction, L_{old} is a previous value of said time fraction, and $f \in (0,1)$ is a parameter controlling a rate of change of said time fraction.

23. The apparatus of claim 20 wherein said processor is configured to select a fast attack mode if all of the following conditions are satisfied:

$L < L_{AMThreshold}$,
$AF < AF_{Idle}$,
the two most recently received packets are good,

wherein $L_{AMThreshold}$ is a threshold controlling the transition to said fast attack mode with respect to L and AF_{Idle} is a threshold controlling the transition to said fast attack mode with respect to AF.

24. The apparatus of claim 20 wherein said processor is configured to select a normal mode if any of said conditions are satisfied:

$L \geq L_{NMThreshold}$,
$AF \geq AF_{Busy}$, or
the most recently received packet is bad,

wherein $L_{NMhreshold}$ is a threshold controlling the transition to said normal mode with respect to L and AF_{Idle} is a threshold controlling the transition to said normal mode with respect to AF.

25. The apparatus of claim 19 wherein when in a fast attack mode of operation said processor is configured to increase said closed loop correction in accordance with an equation:

$$L_{new} = \min(L_{old} + \delta', L_{max}),$$

wherein L_{new} is an updated value of said outer loop correction, L_{old} is a previous value of said outer loop correction, δ' is a step size, and L_{max} is the maximum value that said outer loop correction can attain.

26. The apparatus of claim 19 wherein when in a normal mode of operation said processor is configured to increase said closed loop correction in accordance with an equation:

$$L_{new} = \min(L_{old} + TARGET\_PER \cdot \delta, L_{max}),$$

wherein L_{new} is an updated value of said outer loop correction, L_{old} is a previous value of said outer loop correction, *TARGET_PER* is a packet error rate to be attained, δ' is a step size, and L_{max} is the maximum value that said outer loop correction can attain.

27. The apparatus of claim 14 wherein the processor is configured to:

sum said open loop prediction of signal to noise and interference ratio and said closed loop correction; and

determine said data rate as the highest data rate, a signal to noise ratio of which is below said modified signal to noise ratio.

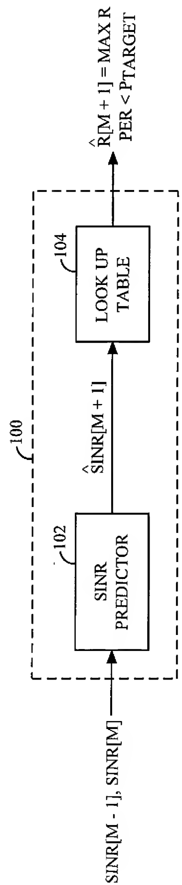
100
102
SINR PREDICTOR
104
LOOK UP TABLE
SINR[M - 1], SINR[M]
$\hat{SINR}$[M + 1]
$\hat{R}$[M + 1] = MAX R
PER < PTARGET

FIG. 1

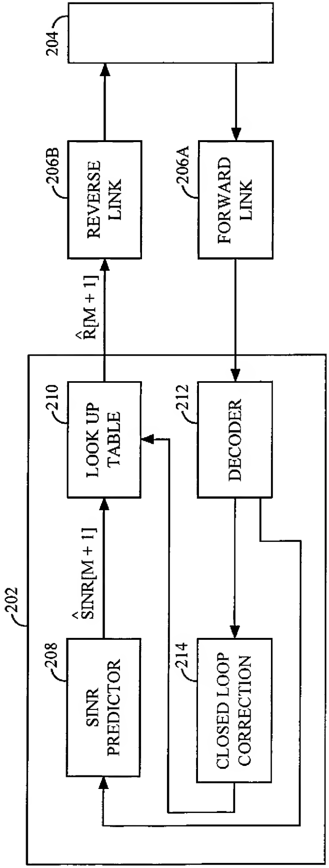
202
204
206B
REVERSE LINK
206A
FORWARD LINK
208
SINR PREDICTOR
210
LOOK UP TABLE
212
DECODER
214
CLOSED LOOP CORRECTION
$\hat{SINR}$[M + 1]
$\hat{R}$[M + 1]

FIG. 2

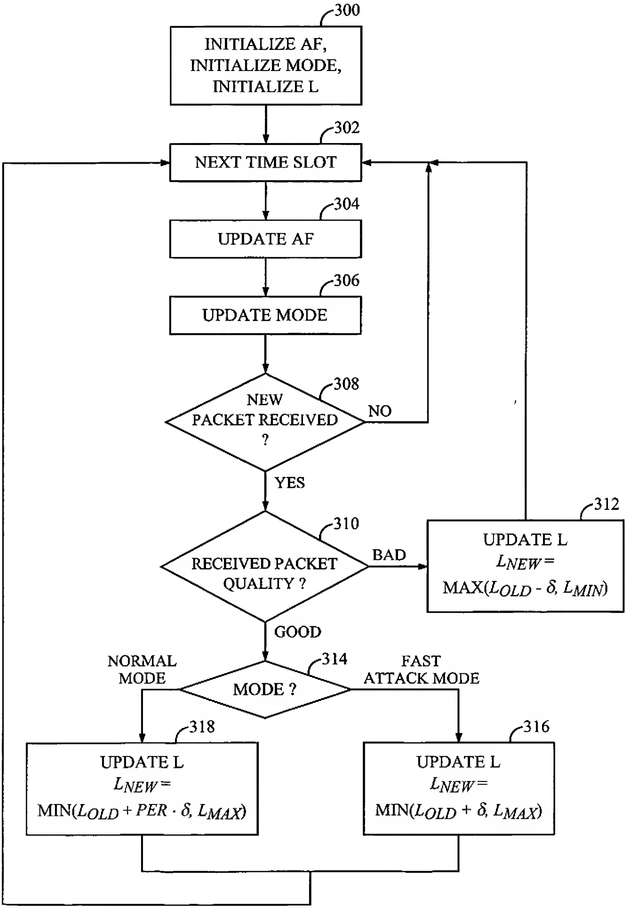
300
INITIALIZE AF,
INITIALIZE MODE,
INITIALIZE L
302
NEXT TIME SLOT
304
UPDATE AF
306
UPDATE MODE
308
NEW
PACKET RECEIVED
?
NO
YES
310
RECEIVED PACKET
QUALITY ?
BAD
312
UPDATE L
L_{NEW} =
MAX(L_{OLD} - δ, L_{MIN})
GOOD
314
MODE ?
NORMAL
MODE
FAST
ATTACK MODE
318
UPDATE L
L_{NEW} =
MIN(L_{OLD} + $PER \cdot \delta$, L_{MAX})
316
UPDATE L
L_{NEW} =
MIN(L_{OLD} + δ, L_{MAX})

FIG. 3

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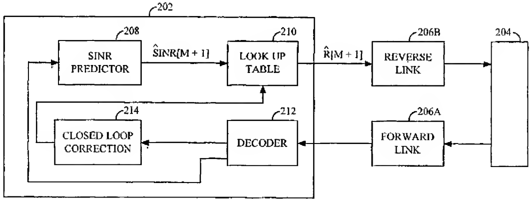
(54) Title: METHOD AND APPARATUS FOR A RATE CONTROL IN A HIGH DATA RATE COMMUNICATION SYSTEM



(57) Abstract: A method and an apparatus for rate control in a high data rate (HDR) communication system are disclosed. An exemplary HDR communication system defines a set of data rates, at which an access point (AP) may send data packets to an access terminal (AT). The data rate is selected to maintain targeted packet error rate (PER). The AT's open loop algorithm measures received signal to interference and noise ratio (SINR) at regular intervals, and uses the information to predict an average SINR over the next packet duration. The AT's closed loop algorithm measures a packet error rate (PER) of the received signal, and uses the PER to calculate a closed loop correction factor. The loop correction factor is added to the SINR value predicted by the open loop, resulting in an adjusted SINR. The AT maintains a look up table, which comprises a set of SINR thresholds that represent a minimum SINR necessary to successfully decode a packet at each data rate. The AT uses the adjusted set of SINR thresholds in the look up table to select the highest data rate, the SINR threshold of which is below the predicted SINR. The AT then requests, over the reverse link, that the AP send the next packet at this data-rate.

WO 01/82521 A3

INTERNATIONAL SEARCH REPORT

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According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	EP 0 899 906 A (LUCENT TECHNOLOGIES INC) 3 March 1999 (1999-03-03) abstract paragraphs '0002!-'0007!,'0014!-'0038!,'0040!,'0041! figures 10-12 ---	1-27
A	US 5 541 955 A (JACOBSMEYER JAY M) 30 July 1996 (1996-07-30) abstract; figures 3,4,23 column 6, line 28 - line 49 column 11, line 45 -column 23, line 29 column 29, line 5 -column 30, line 2 column 34, line 9 -column 38, line 51 -----	1-27

☐ Further documents are listed in the continuation of box C.

☒ Patent family members are listed in annex.

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Authorized officer

Borges, P

INTERNATIONAL SEARCH REPORT

Information on patent family members

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			CA	2244428 A1	24-02-1999
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			WO	9411955 A1	26-05-1994

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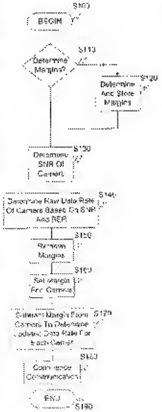
[Continued on next page]

(54) Title: SYSTEMS AND METHODS FOR A MULTICARRIER MODULATION SYSTEM WITH A VARIABLE MARGIN



(57) Abstract: A multicarrier modem has a plurality of carriers over which data is transmitted. By assigning, for example, one or more different margins to the individual carriers, the data rate and impairment immunity can be increased.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

DATA ALLOCATION WITH VARIABLE SNR MARGINS

Background of the Invention

Field of the Invention

[0001] This invention relates to communications technologies. In particular, this invention relates to multicarrier modulation systems having multiple margins.

Description of Related Art

[0002] Multicarrier modulation, or Discrete Multitone Modulation (DMT), is a transmission method that is widely used for communication over difficult media. Multicarrier modulation divides the transmission frequency band into multiple subchannels, i.e., carriers or bins, with each carrier individually modulating a bit or a collection of bits. A transmitter modulates an input data stream containing information bits with one or more carriers, i.e., bins or subchannels, and transmits the modulated information. A receiver demodulates all the carriers in order to recover the transmitted information bits as an output data stream.

[0003] Multicarrier modulation has many advantages over single carrier modulation. These advantages include, for example, a higher immunity to impulse noise, a lower complexity equalization requirement in the presence of multipath, a higher immunity to narrow band interference, a higher data rate and bandwidth flexibility. Multicarrier modulation is being used in many applications to obtain these advantages, as well as for other reasons. These applications include Asymmetric Digital Subscriber Line (ADSL) systems, wireless LAN systems, power line communications systems, and other applications. ITU standards G.992.1 and G.992.2 and the ANSI T1.413 standard specify standard implementations for ADSL transceivers that use multicarrier modulation.

[0004] Discrete multitone modulation transceivers modulate a number of bits on each subchannel, the number of bits depending on the Signal to Noise Ratio (SNR) of that subchannel and the Bit Error Rate (BER) requirement of a link. For example, if the required BER is 1×10^{-7}, i.e., one bit in ten million is received in error on average, and the

SNR of a particular subchannel is 21.5 dB, then that subchannel can modulate 4 bits, since 21.5 dB is the required SNR to transmit 4 QAM bits with a 1×10^{-7} BER. Other subchannels can have a different SNR and therefore may have a different number of bits allocated to them at the same BER. Additional information regarding bit loading can be found in copending U.S. Application Serial No. 09/510,773, incorporated herein by reference in its entirety.

[0005] In many DMT systems, an additional parameter is used to determine the number of bits allocated to each subchannel. This parameter is called the SNR "margin," or simply the "margin." The margin specifies an extra SNR per subchannel, in addition to what is required to maintain the specified BER requirement. As an example, a DMT system with a 6 dB margin would require a 21.5+6=27.5 dB SNR on a subchannel in order to transmit 4 bits on that subchannel with a 1×10^{-7} BER. This is 6 dB more than required by the example in the previous paragraph because now a 6 dB margin is added to the system. Another way of looking at this is that in the example of the previous paragraph, where 4 bits were allocated to a subchannel with 21.5 dB SNR, the margin was 0 dB.

[0006] DMT transceivers use a margin to increase the system's immunity to various types of time varying impairments. Examples of these impairments in DSL systems are: changes in the levels of crosstalk from other transmission systems, impulse noise, temperature changes in the telephone line, or the like. When a DMT system is operating with a positive SNR margin, the noise can change instantaneously by the level of the margin and the system will still maintain the required BER. For example, if the system is operating at a 6 dB margin, e.g., 4 bits are allocated to carriers with 27.5 dB SNR for BER=1×10^{-7}, the crosstalk levels can increase by 6 dB and the system will still be operating at the required 1×10^{-7} BER. Obviously the penalty for this increase in robustness is a decrease in the data rate, since with a 0 dB margin, a subchannel with 27.5 dB SNR can modulate 6 bits at 1×10^{-7} BER.

[0007] Therefore, there is a tradeoff between the robustness of the channel, such as a phone line, and the achievable data rate. The margin can be used to quantify this tradeoff. A higher margin results in a higher level of immunity to changing channel conditions at the expense of the achievable data rate. Likewise, a lower margin results in a higher data rate at the expense of a lower immunity to changing channel conditions

[0008] Current DMT systems allocate a fixed margin to all subchannels. For example, ADSL systems typically use a 6 dB margin on all subchannels carrying data bits. This 6 dB margin is constant on all subchannels and is independent of the type of impairment that the margin is trying to protect against.

SUMMARY OF THE INVENTION

[0009] For simplicity of reference, the systems and methods of this invention will hereinafter refer to the transceivers, or multicarrier modems, generically as modems. One such modem is typically located at a customer premises such as a home or business and is "downstream" from a central office with which it communicates. The other modem is typically located at the central office and is "upstream" from the customer premises. Consistent with industry practice, the modems are often referred to as "ATU-R" ("ADSL transceiver unit, remote," i.e., located at the customer premises) and "ATU-C" ("ADSL transceiver unit, central office," i.e., located at the central office). Each modem includes a transmitter section for transmitting data and a receiver section for receiving data, and is of the discrete multitone type, i.e., the modem transmits data over a multiplicity of subchannels of limited bandwidth. Typically, the upstream or ATU-C modem transmits data to the downstream or ATU-R modem over a first set of subchannels, which are usually the higher-frequency subchannels, and receives data from the downstream or ATU-R modem over a second, usually smaller, set of subchannels, commonly the lower-frequency subchannels.

[0010] For example, in digital subscriber line (DSL) technology, communications over a local subscriber loop between a central office and a subscriber premises is accomplished by modulating the data to be transmitted onto a multiplicity of discrete frequency carriers which are summed together and then transmitted over a subscriber loop. Individually, the carriers form discrete, non-overlapping communication subchannels which are of a limited bandwidth. Collectively, the carriers form what is effectively a broadband communications channel. At the receiver end, the carriers are demodulated and the data recovered.

[0011] DSL systems experience disturbances from other data services on adjacent phone lines, such as, for example, ADSL, HDSL, ISDN, T1, or the like. Additionally, DSL systems may experience disturbances from impulse noise, crosstalk, temperature changes, or the like. These disturbances may commence after the subject DSL service is already

initiated and, since DSL for Internet access in envisioned as a always-on service, the affects of these disturbances should be considered by the subject DSL transceiver. Additionally, the length of the phone line is a type of impairment that varies from one ADSL subscriber to another, i.e. from one ADSL installation to another, and therefore has an effect on the ADSL modem performance.

[0012] The systems and methods of this invention allow the margin in a discrete multitone modulation system to vary depending on a type of impairment. For example, this impairment can be changing over some duration or from one installation to another. Thus, different margins can be assigned to one or more of the carriers in a discrete multitone modulation communication system.

[0013] As noted above, there is a tradeoff between the robustness of the link and the achievable data rate. By setting a higher margin, a higher level of immunity to changing channel conditions is achieved at the expense of the data rate. Similarly, while a lower margin may result in a higher data rate, the immunity to changing channel conditions is reduced.

[0014] However, setting the margin equally for all subchannels at least fails to account for impairments that change over time and how the impairments may have different effects on subchannels at different frequencies. For example, temperature changes and line length effect different frequencies with differing degrees of interference.

[0015] Aspects of the present invention relate to a communications system having a plurality of margins.

[0016] Aspects of the present invention also relates to a method of assigning a plurality of margins to a communications system.

[0017] Aspects of this present invention additionally relate to multicarrier modulation systems and methods for different margins to be assigned to different subchannels to account for varying impairments.

[0018] These and other features and advantages of this invention are described in, or are apparent from, the following detailed description of the embodiments

BRIEF DESCRIPTION OF THE DRAWINGS

[0019] The embodiments of the invention will be described in detail, with reference to the following figures wherein:

[0020] Fig. 1 is a functional block diagram illustrating an exemplary modem according to this invention; and

[0021] Fig. 2 is a flowchart outlining an exemplary method for assigning margins according to this invention.

DETAILED DESCRIPTION OF THE INVENTION

[0022] In an exemplary embodiment of the invention, the margin is set to be different on at least two subchannels in a discrete multitone modulation system. In this exemplary embodiment, subchannels which are expected to incur greater variations in impairment levels are set to have a higher margin, whereas subchannels which are expected to incur lower variations in impairment levels are set to have lower margins. As an example of this embodiment, consider an ADSL transmission system transmitting data over telephone wires and consider the case where the impairment is changing channel conditions due to temperature fluctuations. Since telephone wire is typically made out of copper, the attenuation, i.e., the insertion loss, characteristics will depend on the temperature of the wire. As the temperature of the wire increases, the attenuation, i.e., the insertion loss, will increase. Furthermore, the insertion loss also varies with frequency as the temperature changes. Therefore, as the temperature increases, in addition to an overall increase in insertion loss, the insertion loss at the higher frequencies increases more than the insertion loss at the lower frequencies. Table 1 shows a correlation of frequency versus insertion loss of an exemplary 13,500 ft. 26 AWG line at various frequencies for 70°F and 120°F.

	Frequency (kHz)										
	20	40	100	200	260	300	400	500	600	780	1100
Insertion loss (dB) at 70° F	29.8	36.7	45.2	52.8	57.3	60.2	67.7	74.8	81.7	93.0	110
Insertion loss (dB) at 120° F	31.9	39.6	49.4	57.4	61.8	64.8	72.3	79.3	86.1	97.9	116

TABLE 1: Insertion loss of 13500 ft 26 AWG line versus frequency at 70F and 120F

[0023] From Table 1, it is apparent that the difference in insertion loss from 120°F to 70°F is 2.1 dB at 20 kHz, whereas the difference in insertion loss from 120°F to 70°F is 6 dB at 1100 kHz. For this exemplary embodiment, a higher margin could be allocated to carriers at higher frequencies and a lower margin allocated to carriers at lower frequencies. For example, the carrier at 20 kHz will only need a 2.1 dB margin, because even if the temperature changes from 70°F to 120°F, the insertion loss will only change by 2.1 dB and, as a result, the system bit error rate requirement can still be met after the temperature change. Similarly, the carrier at 1100 kHz will need a 6 dB margin, since as the temperature changes from 70°F to 120°F, the insertion loss will change by 6 dB and, as a result, the system bit error rate requirement will still be satisfied even after the temperature change.

[0024] However, it is to be appreciated that the margin is not allocated to each subchannel in a fixed manner, but rather varies based on the expected change in impairments over time or as impairments vary from one DSL installation to another. However, that does not preclude the possibility that different subchannels can have the same margin assigned to them. For example, a subchannel may have a certain margin assigned based on a particular impairment, while another subchannel may have the same margin assigned based on another impairment. These impairments can include, but are not limited to, changes in the levels of crosstalk from other transmission systems, impulse noise, temperature changes, line length, radio frequency interference and other ingress, or the like. As a result, for example, since certain subchannels are not overly burdened with a common margin, the overall data rate of the system can be increased without sacrificing the robustness of the system.

[0025] For example, and with reference to Table 1, by lowering the margin of the lower carriers from 6 dB to 2.1 dB, the channel data rate has increased. This increase can occur without a loss of the immunity to temperature variations on the line since the lower frequencies are less susceptible to temperature changes than the higher frequencies. In general, the systems and methods of this invention can be adapted to set a margin for any impairment that varies over time, or is installation based, and may, for example, effect different frequencies in different ways.

[0026] As another example, consider crosstalk from another transmission system. If the crosstalking transmission system is known to use only a portion of the frequency spectrum utilized by the discrete multitone modulation system, then the margins can be decreased on the carriers that are known to be outside the frequency spectrum of the crosstalking system. For example, ISDN systems are an example of a crosstalk source for ADSL systems. ISDN systems typically transmit only up to approximately 150 kHz. Thus, for example, employing the teachings of this invention, carriers above 150 kHz can operate at lower margins than carriers below 150 kHz where the ISDN crosstalk is present.

[0027] As another example, the margin in an ADSL system can be varied depending on the length of the telephone wire. Table 2 shows a relationship of insertion loss of an exemplary 9000 ft. 26 AWG line at frequencies for 70°F and 120°F.

	Frequency (kHz)										
	20	40	100	200	260	300	400	500	600	780	1100
Insertion loss (dB) at 70° F	20.0	24.4	30.1	35.2	38.2	40.2	45.1	49.9	54.4	62.0	73.6
Insertion loss (dB) at 120° F	21.4	26.3	32.8	38.2	41.2	43.2	48.2	52.9	57.4	65.3	77.5

TABLE 2: Insertion loss of 9000 ft 26 AWG line versus frequency at 70F and 120F

[0028] Comparing Table 1 and Table 2, it is apparent that an increase in insertion loss as temperature increases depends on the length of the telephone line as well. Thus, on the exemplary 9,000 ft. phone line, a 50°F temperature change results in an average of only 2.8 dB increase in insertion loss. On the 13,500 ft. phone line, a 50°F temperature change resulted in an average of 4.3 dB increase in insertion loss. For this illustrative example, the

margin on the subchannels is varied depending on the length of the phone line. As an example, if the phone line is shorter, e.g., 9,000 ft., the average margin can be decreased on the subchannels by 4.3-2.8=1.5 dB as compared to a longer 13,500 ft. loop without sacrificing immunity to temperature changes on the phone line. This is possible because a shorter phone line will not experience as much of a change in insertion loss due to temperature changes as a longer phone line.

[0029] For this illustrative example, the margin allocated to different subchannels takes into account information about the length of the telephone line. As an example, the insertion loss difference from 70°F to 120°F at 20 kHz is 2.1 dB for the 13,500 ft. line. On the other hand, the insertion loss difference from 70°F to 120°F at 20 kHz is 1.4 dB for the 9,000 ft. line. Therefore, for this exemplary situation, a margin of 2.1 dB would be allocated to the carrier at 20 kHz on a 13,500 ft. line whereas a margin of 1.4 dB would be allocated to the carrier at 20 kHz on the 9,000 ft. line. The immunity to temperature variations on the line would be the same for both the systems operating at 9,000 ft. and 13,500 ft. As a result, the overall system data rate can be increased on shorter lines without sacrificing a loss in robustness.

[0030] Fig. 1 illustrates an exemplary embodiment of a multicarrier modem 100. In particular, the multicarrier modem 100 comprises a controller 10, a memory 20, a discrete multitone modulation system 30, a data rate determiner 40, a signal to noise ratio determiner 50, a margin determiner 60 and a margin storage 70, all interconnected by link 5. The multicarrier modem 100 is also connected to one or more computer or computer-type devices 80 and additional modems (not shown) via communications link 10. For ease of illustration, the multicarrier modem 100 has been illustrated in block diagram format with only the components needed for the exemplary embodiment of this invention. Additional information and further discussion of the operation and structure of an exemplary multicarrier modem can be found in copending U.S. Patent Application Serial No. 09/485,614 entitled "Splitterless Multicarrier Modem," incorporated herein by reference in its entirety.

[0031] While the exemplary embodiment illustrated in Fig. 1 shows the multicarrier modem 100 and various components collocated, it is to be appreciated that the various components of the multicarrier modem can be rearranged and located in whole or in part at

an ATU-R and/or ATU-C. Furthermore, it is to be appreciated, that the components of the multicarrier modem 100 can be located at various locations within a distributed network, such as a POTS network, or other comparable telecommunications network. Thus, it should be appreciated, that the components of the multicarrier modem 100 can be combined into one device or distributed amongst a plurality of devices. As will be appreciated from the following description, and for reasons of computational efficiency, the components of the multicarrier modem can be arranged at any location within a telecommunications network and/or modem without affecting the operation of the system.

[0032] The links 5 and 10 can be a wired or a wireless link or any other known or later developed element(s) that is capable of supplying and communicating electronic data to and from the connected elements. Additionally, the computer device 80, can be, for example, a personal computer or other device. In general, the computer device 80 can be any device that uses a modem to transmit and/or receive data.

[0033] In operation, the multicarrier modem 100 is installed, for example, in a customer premises or in a central office. During this installation, certain fixed quantities such as line length are known and can be stored in the multicarrier modem 100. During an initial installation, or at any subsequent time for which a redetermination in margins is appropriate, for example, based on an increased bit error rate, changes in the signal to noise ratio, seasonal changes, or the like, the controller 10, in cooperation with the memory 20, the discrete multitone modulation system 30 and the margin determiner 60 can determine and store margins. For example, as illustrated above in exemplary Tables 1 and 2, margins can be determined for temperature fluctuations and the length of the wire line based on, for example, the actual installation and historical data. Furthermore, routines can be established by the margin determiner 60 to evaluate and compile statistical information relating to one or more carriers. For example, this statistical information can be compiled during modem idle times in response to impairments seen on the one or more carriers. This statistical information can then be used to determine appropriate margins for one or more carriers.

[0034] Alternatively the modem may measure the noise on the line during idle times and determine that a particular type of crosstalker, e.g., another ADSL or HDSL modem, is present. Since the spectral content of these types of crosstalkers are known, this information can be used to determine the margin. For example, if the crosstalker is an ATU-R ADSL

modem then it is known that ATU-R ADSL modems transmit approximately in the 20-130 kHz range. This information can be used to determine the margin for the carriers in the 20-130 kHz frequency range.

[0035] Alternatively, a predetermined set of margins, for example, for known impairments, can be downloaded from, for example, a central office modem or other location within a communications network. The determined and/or downloaded margins are then stored in the margin storage 70. Similarly, groups of margins can be stored based on, for example, geographic information, seasonal information, line length information, or the like.

[0036] During training of the multicarrier modem 100, the SNR determiner 50, in cooperation with the controller 10, the memory 20, and the DMT system 30, determines the signal to noise ratio of the carriers. Knowing the signal to noise ratio of the carriers, the data rate determiner 40 determines the raw data rate of the carriers based on the signal to noise ratio and the bit error rate. This raw data rate reflects the data rate of carriers with no margin.

[0037] Generally, the bit error rate is set in advance, for example, by the manufacturer. Additionally, the data rate is generally governed by a range that is, for example, guaranteed as a maximum, by a DSL provider. Therefore, based on the set bit error rate, the signal to noise ratio for a known quantity of bits can be determined.

[0038] Knowing the signal to noise ratio, the margins for the carriers can be set, for example, based on one or more, or a combination of, entered criteria or determined criteria. For example, an entered criteria can be based on the loop length. A determined criteria can be, for example, based on standard temperature variance information that can, for example, be downloaded from the service provider. Alternatively, for example, the margins can be set based on historical data that relates to, for example, impairments on the line. In general, the margins can be set such that a balance between the data rate and the impairment immunity is maximized.

[0039] Having retrieved the margins for one or more of the carriers, the margins are set in the DMT system 30. The margins can then be subtracted from the carrier to determine an

updated data rate for each carrier. Having set the margins, and knowing the data rate, the DMT system can then commence communication over the communications link 10.

[0040] Fig. 2 illustrates an exemplary method of assigning margins to carriers according to an exemplary embodiment of this invention. In particular, control begins in step S100 and continues to step S110. In step S110, a determination is made whether margins are to be determined. If margins are to be determined, control continues to step S120. Otherwise, control jumps to step S130.

[0041] In step S120, the margins are determined and stored. Control then continues to step S130.

[0042] In step S130, the signal to noise ratio of the carriers are determined. Next, in step S140, the raw data rate of the carriers is determined based on the signal to noise ratio and the bit error rate. Next, in step S150, the margins for the carriers are retrieved. Control then continues to step S160.

[0043] In step S160, the margins for the carriers are set. Next, in step S170, the margins are subtracted from the carriers to determine an updated data rate for each carrier. Control then continues to step S180.

[0044] In step S180, communications commence. Control then continues to step S190 where the control sequence ends.

[0045] However, it is to be appreciated that the steps in Fig. 2 need not occur in the order illustrated. For example, at any point in time there could be an option to re-determine the margins. Similarly, based on, for example, the time of the day, day, location, error rate, service provider directive, a change in the quality of service requirement, or the like, the margins could be adjusted. Alternatively, at any time, updated margins could be downloaded and stored in the margin storage. Alternatively, if it known that margins will be incorporated in the determination of the data rate, step S140 could be bypassed since it is known that the raw data rate will not be used

[0046] Furthermore, the systems and methods of this invention can also apply to any multicarrier modulation based communication system including wireless LANs, such as wireless LAN 802.11 and ETSI Hyperlan standards, wireless access systems, home and access power-line communication systems, or the like.

[0047] As illustrated in Fig. 1, the multicarrier modem and related components can be implemented either on a DSL modem, or a separate program general purpose computer having a communications device. However, the multicarrier modem can also be implemented in a special purpose computer, a programmed microprocessor or microcontroller and peripheral integrated circuit element, and ASIC or other integrated circuit, a digital signal processor, a hardwired or electronic or logic circuit such as a discrete element circuit, a programmable logic device such as a PLD, PLA, FPGA, PAL, or the like, and associated communications equipment. In general, any device capable of implementing a finite state machine that is in turn capable of implementing the flowchart illustrated in Fig. 2 can be used to implement the multicarrier modem 100 according to this invention.

[0048] Furthermore, a disclosed method may be readily implemented in software using object or object-oriented software development environment that provides portable source code that can be used on a variety of computers, workstations, or modem hardware platforms. Alternatively, the disclosed modem may be implemented partially or fully in hardware using standard logic circuits or a VLSI design. Other software or hardware can be used to implement the systems in accordance with this invention depending on the speed and/or efficiency requirements of the systems, the particular function, and the particular software or hardware systems or microprocessor or microcomputer systems being utilized. The multicarrier modem illustrated herein, however, can be readily implemented in hardware and/or software using any known or later developed systems or structures, devices and/or software by those of ordinary skill in the applicable art from the functional description provided herein and with a general basic knowledge of the computer and telecommunications arts.

[0049] Moreover, the disclosed methods can be readily implemented as software executed on a programmed general purpose computer, a special purpose computer, a microprocessor and associated communications equipment, or the like. In these instances, the methods and systems of this invention can be implemented as a program embedded on a modem, such as a DSL modem, or the like. The multicarrier modem can also be implemented by physically incorporating the system and method in a software and/or hardware system, such as a hardware and software system of a modem, such as an ADSL modem, or the like.

[0050] It is, therefore, apparent that there has been provided in accordance with the present invention, systems and methods for assigning margins to carriers. While this invention has been described in conjunction with a number of embodiments, it is evident that many alternatives, modifications and variations would be or are apparent to those of ordinary skill in the applicable art. Accordingly, Applicants intend to embrace all such alternatives, modifications, equivalents and variations that are within the spirit and the scope of this invention.

What is Claimed is:

1. A multicarrier modulation communication system comprising:
a plurality of subchannels; and
a plurality of margins.

2. The system of claim 1, wherein the plurality of margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

3. The system of claim 1, wherein the plurality of margins are at least one of an average margin and a subchannel specific margin.

4. The system of claim 3, wherein the average margin is applied equally to a portion of the plurality of subchannels.

5. The system of claim 1, further comprising a margin determiner that determines at least one margin.

6. The system of claim 1, further comprising a margin storage device that stores at least one margin.

7. A multicarrier modulation communication system comprising:
a plurality of subchannels; and
a plurality of margins, wherein one of the plurality of margins is assigned to at least one of the plurality of subchannels.

8. The system of claim 7, wherein the plurality of margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

9. The system of claim 7, wherein the plurality of margins are at least one of an average margin and a subchannel specific margin.

10. The system of claim 9, wherein the average margin is applied equally to a portion of the plurality of subchannels.

11. The system of claim 7, further comprising a margin determiner that determines at least one margin.

12. The system of claim 7, further comprising a margin storage device that stores at least one margin.

13. A multicarrier modulation communication system communicating on a wire line over a plurality of subchannels, wherein at least one margin based on a length of the wire line is assigned to at least one of the plurality of subchannels.

14. The system of claim 13, wherein the at least one margin is based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

15. The system of claim 13, wherein the at least one margin is based on at least one of an average margin and a subchannel specific margin.

16. The system of claim 15, wherein the average margin is applied equally to a portion of the plurality of subchannels.

17 The system of claim 13, further comprising a margin determiner that determines at least one margin.

18. The system of claim 13, further comprising a margin storage device that stores at least one margin.

19 A multicarrier modulation communication system having a plurality of subchannels, wherein at least two subchannels have a different margin.

20 The system of claim 19, wherein the margin is based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

21. The system of claim 19, wherein the margins are at least one of an average margin and a subchannel specific margin.

22. The system of claim 21, wherein the average margin is applied equally to a portion of the at least two subchannels.

23. An information storage media comprising margin information for a multicarrier modulation system having a plurality of subchannels, wherein at least two subchannels have a different margin.

24. A method of enhancing multicarrier modulation communication over a plurality of subchannels comprising communicating over the plurality of subchannels using at least two different margins.

25. The method of claim 24, wherein the at least two margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

26. The method of claim 24, wherein the at least two margins are at least one of an average margin and a subchannel specific margin.

27. The method of claim 26, wherein the average margin is applied equally to a portion of the plurality of subchannels.

28. The method of claim 24, further comprising determining at least one margin.

29. The method of claim 24, further comprising storing at least one margin.

30. A method for multicarrier modulation communication over a plurality of subchannels comprising:

selecting a first number of the subchannels;

assigning a first margin to the first number of the subchannels;

selecting a second number of the subchannels; and

assigning a second margin to the second number of subchannels, wherein the first margin and the second margin are different.

31. The method of claim 30, wherein the margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

32. The method of claim 30, wherein the margins are at least one of an average margin and a subchannel specific margin.

33. The method of claim 32 wherein the average margin is applied equally to a portion of either the first or second number of subchannels.

34. The method of claim 30, further comprising a margin determiner that determines at least one margin.

35. The method of claim 30, further comprising a margin storage device that stores at least one margin.

36. A method for multicarrier modulation communication over a wire line using a plurality of subchannels, wherein at least one margin based on a length of the wire line is assigned to at least one of the plurality of subchannels.

37. The method of claim 36, wherein the at least one margin is based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

38. The method of claim 36, wherein the at least one margin is at least one of an average margin and a subchannel specific margin.

39. The method of claim 38, wherein the average margin is applied equally to a portion of the plurality of subchannels.

40. The method of claim 36, further comprising a margin determiner that determines at least one margin.

41. The method of claim 36, further comprising a margin storage device that stores at least one margin.

42. A method for communicating in a multicarrier modulation communications environment having at least two subchannels, wherein at least two of the at least two subchannels have a different margin.

43 The method of claim 42, wherein the margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

44 The method of claim 42, wherein the margins are at least one of an average margin and a subchannel specific margin.

45. The method of claim 42, wherein the average margin is applied equally to a portion of the at least two subchannels.

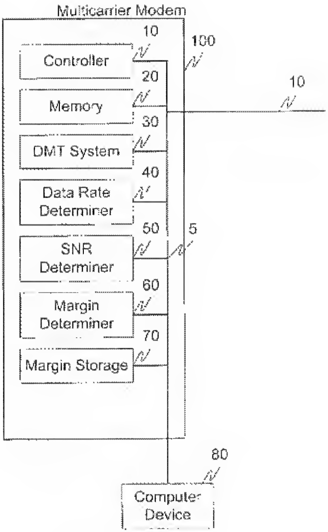
Multicarrier Modem
10
100
Controller
20
10
Memory
30
DMT System
40
Data Rate
Determiner
50
5
SNR
Determiner
60
Margin
Determiner
70
Margin Storage
80
Computer
Device

Fig. 1

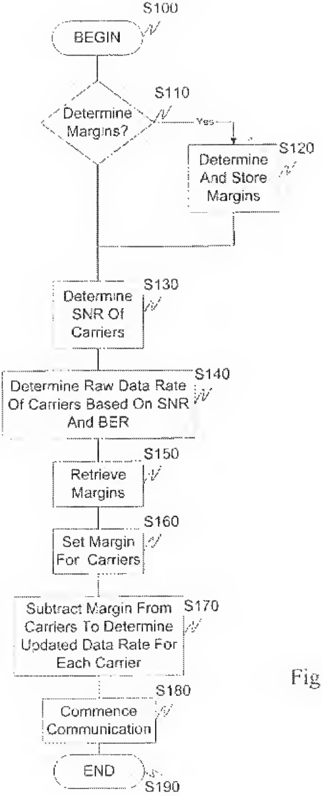
S100
BEGIN
S110
Determine Margins?
Yes
S120
Determine And Store Margins
S130
Determine SNR Of Carriers
S140
Determine Raw Data Rate Of Carriers Based On SNR And BER
S150
Retrieve Margins
S160
Set Margin For Carriers
S170
Subtract Margin From Carriers To Determine Updated Data Rate For Each Carrier
S180
Commence Communication
END
S190

Fig. 2

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 01/12555

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L27/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, PAJ, WPI Data, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 99 20027 A (AWARE) 22 April 1999 (1999-04-22) page 11, line 18 - line 23 page 11, line 26	19-29
X	COLIN ET AL.: "QoS considerations for DMT-based ADSL and VDSL systems" IEEE INTERNATIONAL CONFERENCE ON ACOUSTICS, SPEECH AND SIGNAL PROCESSING, 12 - 15 May 1998, pages 3437-3440, XP000951196 New York, US ISBN: 0-7803-4429-4 page 3438, left-hand column, paragraph 4	19-29
X	EP 0 918 422 A (MOTOROLA INC) 26 May 1999 (1999-05-26) column 4, line 6 - line 17	19-29

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[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
26 September 2001	09/10/2001

Name and mailing address of the ISA	Authorized officer
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INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 01/12555

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 5 822 374 A (LEVIN) 13 October 1998 (1998-10-13) figure 3	19-29
X	EP 0 955 744 A (NEC) 10 November 1999 (1999-11-10) page 25, line 30 - line 31 page 26, line 15 - line 19	19-29
A	WO 98 57472 A (AWARE) 17 December 1998 (1998-12-17) page 5, line 10 - line 11	19,23,24
A	"Spectrum management for loop transmission systems" ANSI DRAFT , 'Online' 1999, pages 1-110, XP002178517 Retrieved from the Internet: <URL:http://ftp.tiaonline.org/tr-41/TR419/WORKING/8-99/-046%20T1E1.pdf> 'retrieved on 2001-09-26! paragraph '4.3.5.1!	20,25

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

Continuation of Box I.2

Claims Nos.: 1-18, 30-45

In view of the large number of independent claims, and also of their wording, which renders it difficult, if not impossible, to determine the matter for which protection is sought, the present application fails to comply with the clarity and conciseness requirements of Article 6 PCT (see also Rule 6.1(a) PCT) to such an extent that a meaningful search is impossible. Consequently, the search has been carried out for those parts of the application which do appear to be clear and concise, namely claims 19-29. These claims define the use of at least two different margins. This feature is essential, as is specified in paratraph '0012! of the description.

The applicant's attention is drawn to the fact that claims, or parts of claims, relating to inventions in respect of which no international search report has been established need not be the subject of an international preliminary examination (Rule 66.1(e) PCT). The applicant is advised that the EPO policy when acting as an International Preliminary Examining Authority is normally not to carry out a preliminary examination on matter which has not been searched. This is the case irrespective of whether or not the claims are amended following receipt of the search report or during any Chapter II procedure.

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US 01/12555

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
WO 9920027	A	22-04-1999	AU	1077699 A	03-05-1999
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			EP	1013043 A1	28-06-2000
			WO	9857472 A1	17-12-1998

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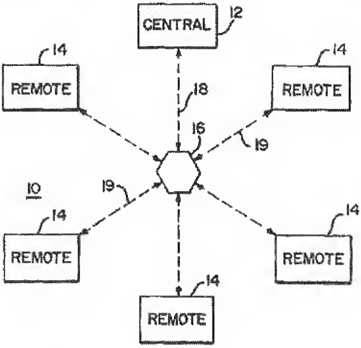
(54) Title: APPARATUS FOR EXCHANGING DATA BETWEEN A CENTRAL STATION AND A PLURALITY OF WIRELESS REMOTE STATIONS



(57) Abstract

An apparatus is provided for exchanging data between a central station (12) and a plurality of wireless remote stations (14) on a time division multiple access communication channel. The apparatus includes means for receiving access requests from remote stations of the plurality of remote stations (14) during a first time interval under a contention based protocol and a non-contention based protocol and means for polling for data transfers during a second time period remote stations (14) of the plurality of remote stations providing access requests under non-contention based protocols during the first time period.

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APPARATUS FOR EXCHANGING DATA BETWEEN A CENTRAL STATION AND A PLURALITY OF WIRELESS REMOTE STATIONS

Background of the Invention

This invention generally relates to the field of data communications networks. More particularly, this invention pertains to a multiple access protocol for a data communications network having a number of users exchanging data between individual remote stations to a central station over a single optical infrared channel.

A multipoint digital communications network typically consists of a number of remote stations which communicate with a central station over one or more two-way communications channels. For example, personal computers are typically connected to a wide variety of peripherals or other computers via wire cables, i.e., a hard-wired communications link. Moreover, local area networks (LAN's) are often used to integrate remote terminals that are located at the same site. Depending upon the number of users, distance between terminals, number of peripherals, frequency of system reconfiguration, portability of the remote stations, etc., the hard-wired cable system may not be practical for a given application. Hence, various wireless communication technologies have been employed, particularly when

a system includes a large number of users and/or portable, hand-held computer devices.

Among the more common wireless technologies are narrow-band radio frequency (RF) systems, spread spectrum RF, ultrasonic, and infrared optical. Radio frequency systems are often significantly degraded by electromagnetic noise and interference, as well as by large signal amplitude variations and multipath interference. Moreover, RF systems are typically subject to governmental licensing and regulation. Alternative wireless systems employing ultrasonic sound waves experience severe problems with the complete loss of signals due to nulls in the transmission path.

Optical-infrared communications, however, is not affected by electromagnetic interference, and is much less susceptible to multipath interference. Furthermore, optical systems are inherently secure (since the infrared light does not penetrate walls), have no known health or safety effects, and are not subject to F.C.C. regulation. Moreover, infrared transceivers draw relatively low currents, which is particularly important with respect to hand-held battery-powered portable computers. Thus, the use of infrared light as the wireless medium is well suited to such applications.

In order for the remote stations to communicate with the central station, the remote stations must be able to gain access to the commonly-shared communications channel using some type of multiple-access signalling or control protocol. As used in the data communications field, a "protocol" is a formal set of rules governing the format and control of inputs and outputs

between two communicating devices in order to ensure the orderly transfer of information. Typical multiple-access protocols may be categorized into two broad classes: contention-based protocols (i.e., random access), and noncontention-based protocols, (i.e., scheduled access). Contention-based protocols are characterized in that any remote user with a data message can contend for the channel by transmitting its data message immediately in an on-demand fashion, taking the chance that no other remote stations will transmit at the same time and thus collide with it. When a collision occurs, the data message is seldom received correctly, if at all. Since there is no coordination between contending remote stations, the number of collisions dramatically increases as the number of users increase, or as the channel load increases. Hence, contention-based protocols are not suitable for many data communications applications.

Noncontention-based protocols are characterized in that they provide the necessary coordination between the remote stations to ensure that no two remote stations transmit at the same time to contend for the channel. In other words, the users in a noncontention system take turns accessing the network in an orderly fashion such that collisions between users are avoided. Noncontention channel access is usually implemented using [illegible]me type of polling technique, wherein the central station se[illegible] a control message or synchronization signal to the remote stations as an indication for the remote to respond by transmitting data on the channel.

Using the well-known "explicit polling" technique, the central controller sends a polling signal to each remote station.

individually, to inquire if the remote has any information to send. A "poll list" of remote station addresses is used by the central controller to determine when a remote station is to be polled. If the polled remote station doesn't have a data message to send over the channel, the central controller goes on to poll the next remote. If the remote station does have a message to send, the data message is immediately transmitted over the channel in response to the poll. As used herein, the term "polling" includes the second-half of the procedure, wherein the polled stations return a message. Explicit polling has traditionally been considered rather inefficient, since each remote station has to wait for its individualized poll, establish bit and character synchronization, and then transmit its data message in response to the poll. Hence, a significant portion of the overall channel capacity is consumed by the polling signals themselves.

Another noncontention-based multiple-access protocol is referred to as "implicit polling." Under the implicit polling technique, each timing cycle on the channel is divided into a number of time slots, and a specific time slot within each cycle is reserved for a particular remote station. Each remote station, which is synchronized in time with the central station, is implicitly granted access to the channel during its individual time slot. In other words, the channel access is controlled by reserving time slots for each remote station to transmit, rather than being controlled by explicit polling signals from the central station.

In multipoint data communications networks using the implicit polling protocol, a fixed transmission time slot is reserved for each remote station in the network. Each time slot must be of a sufficient length to contain an entire data message packet. Hence, the channel is efficiently utilized only if each remote station has a data message to send during each cycle. If, however, only a few of the remote stations have messages to send during each cycle, then the channel remains idle during the preassigned time slots allocated to these non-responsive remote stations. When only a fraction of the remote users have data messages to send, an enormous amount of channel capacity is wasted in the empty time slots of an implicit polling system.

One advance over the prior art was provided by U.S. Patent No. 5,297,144 ("the '144 patent") assigned to the same assignee as the present invention. The '144 patent avoids some of the disadvantages of explicit and implicit polling by periodically allowing remote data stations to register a need for a data transmission with the central station under an implicit polling format. Registration is allowed under the '144 patent whenever the central station transmitted a reservation sync ("RS") frame. Contention was avoided following the RS frame by assigning different delay periods to each remote terminal for transmission of an access request following the RS frame.

Under the '144 patent a relatively fixed time period was allocated for the RS frame and access requests ("the reservation request period"). Following the reservation request period a second, variable length, time period is allowed for

polling the remote station and for transfer of data from the requesting remotes ("the polled data transfer period").

While the '144 patent ostensibly reduces power consumption within the remote stations through use of "sleep modes" such power savings is not practical where a remote station is to receive data from the control station. Under the teachings of the '144 patent, a remote station remains active during the polled data transfer mode (does not enter the sleep mode) only long enough to be polled and transfer data. Since data transfer from the central station to the remote station occurs at the end of the polled data transfer mode, and since the polled data transfer mode is of variable length, the sleep mode of the '144 patent cannot be used where data is to be transferred from the central unit to remote stations.

The '144 patent also allows for the addition of new remote stations to the relatively fixes reservation request period through the use of a "membership acquisition period". The membership acquisition period is a multiframe structure within the superframes after the polled data transfer period ("PDTP") wherein the central station accepts new remote stations (inserts new slots within the reservation request period). The membership acquisition period is a fixed time period within the superframe wherein a new remote station (or group of new remote stations) may seek to gain access to the communication system.

While the '144 patent has provided a significant advance over the prior art, the '144 patent still fails to provide a convenient method of coping within rapid membership changes. The '144 patent also fails to address the issue of

power savings where downlink transmissions must occur between the central station and remote stations. Because of the importance of power savings in portable data devices linked to a central station, a need exists for a method and means of remote station power control under dynamic loading conditions involving the two-way exchange of data between remote stations and the central station.

Summary of the Invention

An apparatus is provided for exchanging data between a central station and a plurality of wireless remote stations on a time divided communication channel. The apparatus includes means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol and means for polling during a second time period remote stations of the plurality of remote stations providing access requests under non-contention based protocols during the first time period.

The apparatus also allows remote stations to exchange data directly. Such direct exchange is possible where the central station acts to coordinate such exchanges while deferring the enablement of other users which may interfere on the communication channel.

Another aspect of the invention provides a second time period where data may be transferred from the central station to individual remote stations. A structure for broadcasting common

information to all remote stations is also provided within the second time frame.

Brief Description of the Drawings

The features of the present invention which are believed to be novel are set forth with particularity in the appended claims. The invention itself, however, together with further objects and advantages thereof, may best be understood with reference to the following description when taken in conjunction with the accompanying drawings, in which:

Figure 1 is a general block diagram of the wireless data communications network according to the present invention;

Figure 2 is a pictorial representation of the channel frame format utilized in the multiple-access signalling protocol of the present invention;

Figure 3 is a timing cycle diagram illustrating the two-stage reservation-based polling protocol and data exchange system of the present invention;

Figure 4A-C provides a summary of network control function by frame type in accordance with the invention along with a description of frame content within individual fields of the frame;

Figure 5 depicts a slot arrangement used within the request period in accordance with the invention;

Figure 6 is a timing cycle diagram similar to that of Figure 3 illustrating slot usage.

Figure 7 is a timing cycle diagram similar to that of Figure 6, wherein acknowledgement signals are returned to the remote stations after each data message;

Figures 8a and 8b are timing diagrams representing the power consumption of the remote station receiver and transmitter, respectively, during the reservation-based polling protocol timing cycle and data exchange of Figure 7;

Figure 9 is a detailed block diagram of one of the remote stations of the data communications network shown in Figure 1; and

Figure 10 is a detailed block diagram of the central station of the data communications network of Figure 1.

Detailed Description of the Preferred Embodiments

The solution to the problem of power savings in a dynamically loaded system requiring the two-way exchange of data between remote stations and a central station lies, conceptually, in mixing contention and non-contention based access protocols and in mapping a data transfer period into uplink and downlink epochs. The prior art has taught that either contention based protocols or non-contention based protocols may be used within access periods gaining entry to a multiple access system. Under the invention, it has been determined that an unexpected increase in efficiency may be achieved by using non-contention access protocols for remote stations requiring frequent data exchanges and contention access protocols to remote stations with less frequent data exchanges.

Mapping of data transfer periods, on the other hand, improves efficiency (reduces power consumption) by allowing an indicia of epoch locations to be transferred to remote stations at predetermined intervals. The indicia of epoch location may then be used by the remote stations to deactivate unnecessary power consuming devices during periods of inactivity.

Referring now to Figure 1, a general block diagram of a wireless multipoint data communications system 10 is shown. The system comprises a central station 12 and a number of remote stations 14. The central station 12 may be a stand-alone data processing and control entity or may be an access point (AP) used in conjunction with other data processors and systems over a larger hard-wired network.

Central station 12 communicates with remote stations 14 through an optical infrared transceiver 16 coupled to the central station via a hard-wired link 18. Each of the remote stations 14 includes an optical infrared transceiver which communicates with the central station by sending and receiving data messages over an infrared link 19. Depending upon the type of network, the central station may utilize the data messages itself, or route the data messages on to a different station in a local area network.

In the preferred embodiment, each of the remote stations is a portable, hand-held, battery-powered computer having an integrated infrared transceiver, as will be described in detail below. The remote stations also include a keypad for data input, and a display for data output. Although the present invention is particularly adapted for two-way communications over

a single-frequency infrared channel transmitting bursts of data packets in the half-duplex mode of operation, the present invention can also be used in full-duplex operation as well as half-duplex operation over single-frequency or split-frequency channels. In the preferred embodiment, infrared link 19 has a 4 Megabit data rate using Return To Zero with Bit Insertion (RZBI) encoding scheme. However, the present invention is not limited for use with only wireless links or the particular type of channel or data communications scheme shown here.

Figure 2 illustrates the specific channel frame format 20 used under the protocol for all information transfer and supervisory commands. The frame format of the invention basically follows the High-level Data Link Control (HDLC) data communications line protocol specification of the CCITT, or the Synchronous Data Link Control (SDLC) protocol specified by IBM. Hence, the published detailed specifications for the HDLC or SDLC protocols may be referred to for a further understanding of the common subject matter.

As shown in Figure 2, each frame is subdivided into a number of individual fields, wherein each field is comprised of a number of 8-bit bytes. The following paragraphs describe channel frame format 20:

<u>Preamble (PRE) 22</u>: This field is a 3-byte field whose purpose is to provide a means of establishing bit synchronization of the receiver with the received signal including the clock recovery signal. The value of the preamble is typically chosen to have a high content of transitions (e.g., "FFFFFF" because in RZBI encoding each "1" bit provides a high-low transition).

<u>Start Delimiter (SD) 24</u>: The purpose of the SD frame is to provide byte synchronization within the receiver. The 8 contiguous bits of the pattern provide a clear indication of the boundary between the "1" bits of the PRE and the bits of the SD. It is a unique "illegal" data structure because the bit insertion of the modulation scheme prevents this number of contiguous zero bits from occurring within the data (anyplace between the SD and ED fields).

<u>Destination Identifier (DID) 26</u>: This field contains the 2-byte address of the station to which the frame is being sent. In other words, in a polling frame, the DID field of a frame transmitted to a remote station first identifies the particular remote station being polled by the central station and then the DID field of a return frame identifies the central station as the destination for the data message being returned by the remote station. Each of the stations is assigned a unique identification code, or address. The remote stations typically receive a new DID address each time the remote station registers with the network 10. However, a dynamic address determination procedure could also be used. In the preferred embodiment, the addresses of remote stations (non-controller stations) begin with hex and increase to the maximum amount of remote stations allowed in the network (e.g., 7FFF hexadecimal). Controller stations (e.g., central station 12) may be assigned other numerical values (e.g., 8000-EEED hexadecimal). A value of FFFF hex in this field denotes a broadcast frame, which would be received by all stations.

Source identifier (SID) 28: This field is the 2-byte address of the station sending the frame. To ensure the integrity of the data being transmitted, both the destination and source addresses are included within each frame.

Type of Field (TYP) 30: The 1-byte field indicates to the receiver how to interpret the frames contents and in effect provides a control function. A summary of the possible types of frames are as follows: RSYNC, MRSYNC, RegRTS, RTS, FORF, DSYNC, EDSYNC, RegCTS, CTS, DATA, MDATA, and ACK. The meaning and content of the types of frames listed may be best understood by reference to FIGS. 4A-C. The use of the frames may be best understood by reference to subsequent sections.

Control Flags: This is a 1-byte control field containing bit-mapped flags, primarily used for supervising commands. In the preferred embodiment, control field 32 includes priority flags and retransmissions flags, which will be described below.

Information (INFO) 34: This is a variable length field used for transferring data. The INFO field 34 is also used in conjunction with certain types of frames (e.g., RSYNC, MRSYNC, DYSNC, and EDSYNC) as a repository for an indicia of epoch location (e.g., the location of upward data transfer period (upward period), broadcast period and downward data transfer period (downward period) within the overall data exchange period (data period)).

Frame Check Sequence (FCS) 36: This 4-byte field is used to detect bit errors which may occur during transmission. In the present embodiment, a 32-bit cyclic redundancy check (CRC)

algorithm is used to detect errors in fields 26, 28, 30, 32, and 34.

End Delimiter (ED) 38 and Postamble (Post) 40: The purpose of the ED 38 is to allow the receiver to detect an end of frame. The purpose of the POST 40 is to prevent the receiver from mistaking an ED/POST combination for an SD/DID combination in that the hexadecimal value of CEEEE could be an invalid DID.

Figure 3 illustrates a repeating frame structure (superframe) used by the system 10 to exchange information between the central station 12 and the remote station 12. Each frame making up the superframe has the frame format described above.

Superframes are not always of the same temporal length. The superframe, in turn, may be divided into a variable length period used for receipt of access requests (request period) 50 and a variable length field used for data exchanges (data period) 51.

The central station 12 identifies the beginning of the superframe to the remote stations 14 by transmission of a request synchronization (RSYNC) frame or a mandatory request synchronization (MRSYNC) frame 52. (The RSYNC frame requires only those remote stations 14 desiring access to respond while the MRSYNC requires all remote stations 14 to respond.) The remote stations 14 identify the RSYNC or MRSYNC frames by reference to the type field of the frame (FIG. 4A-C). In addition to identifying the beginning of the superframe, the RSYNC or MRSYNC frame 52 provides information within the INFO field 34 (FIG. 4A) relative to the number and type of slots (slots using a non-contention

based access protocol or a contention based access protocol) within the request period 50. The slot information is used by the remote stations to facilitate system access (to be explained later) or to power-down during the request period 50 if the remote station 14 does not need access to the network 10.

Following the request period 50, the network 10 enters a data period 51. The central station 12 identifies the beginning of the data period 51 to the remote station 14 by transmission of a data descriptor frame 53 (e.g., a data synchronization (DSYNC) or extended data synchronization (EDSYNC) frame). Contained within the INFO field 34 (FIG. 4A) of the DSYNC or EDSYNC frame 53 is temporal information relative to the length of each subsection of the data period 51. The temporal information, as above, is used by the remote stations 14 to reduce a duty cycle of activation by powering-down during appropriate portions of the data period 51.

In accordance with an embodiment of the invention, the slots of the request period are divided into two groups where a first group of slots allows for random access under a contention based protocol (contention slots) and a second group of slots allows for access under a non-contention protocol (reserved slots) (e.g., under an implied polling protocol). Under the invention, the number of contention slots may be constant or may vary based upon an estimate of the number of unregistered remote stations within the service coverage area of the network 10. The number of reserved slots, on the other hand, is adjusted based upon loading. When a remote station 14 is first activated the remote station 14 is granted access to the network 10 under a

two-step process. First the remote station 14 transmits an access request within a contention slot. The central station 12 upon receipt of the access request within the contention slot then, as a second step, assigns the remote station 14 to a non-contention slot before finally granting access.

The remote station 14 first monitors for a RSYNC or MRSYNC frame 52. Since the remote station 14 does not yet have a reserved slot, the remote station 14 must access the network 10 through a contention slot. The remote station 14 identifies contention slots by examining the contents of the INFO field 34 of the RSYNC or MRSYNC frame 52. Contained inter alia within the INFO field 34 of the RSYNC or MRSYNC frame (FIG. 4A) is the total number of slots in the request period and the total number of reserved slots. By knowing the location of the reserved and contention slots relative to the RSYNC or MRSYNC frame (e.g., the non-contention slots may immediately follow the RSYNC or MRSYNC frame), the remote station 14 can determine the location of the contention slots. Access may then be secured through a randomly selected contention slot.

By way of example, FIG 5 depicts a request period having 10 slots. If the reserved slots were designated as being slots 1-7, then slots 8-10 would be the contention slots. An INFO field 34 of a RSYNC or MRSYNC frame 52 in such a case would indicate a total slot number of 10 and a total reserved slot number of 8. Using known methods, the remote station would then randomly generate a number in the range of 1 to 3 and add the randomly selected number to 8 for a final determination of the contention slot to be used in requesting access.

In requesting access to the network 10, the remote station 14 sends a registration request to send (RegRTS) frame (FIG. 4B) within the selected contention slot. The INFO field 34 of the RegRTS frame contains a 48 bit address of the requesting remote station 14 along with coding within the type field that the frame is a RegRTS frame.

Upon receipt of the RegRTS from the remote station 14 by the central station 12, the central station 12 verifies by reference to a memory (not shown) that the address of the remote station 14 that the station is authorized to access the network 10 and that the remote station 14 has a software version compatible with the network 10. Upon verifying that the remote station 14 is an authorized user and is compatible with the network 10, the central station 12 issues a local identifier in favor of the remote station 14. The central station 12, on the other hand, does not immediately transmit the local identification to the remote station. Under the invention the central station waits until the next downward portion of the data period 51 before transmitting the identifier to the requesting remote station 14.

Contained within the local identifier is an identifier of a reserved slot of the request period 50 allocated for use by the remote station 14. The central station 12 may create a reserved slot for the remote station 14 by expanding the length of the request period to 11 slots or may assign the remote station 14 to an unoccupied slot of reserved slots 1-8 (FIG. 5).

Likewise, the central station 12 may de-allocate a slot previously reserved for use by other remote stations 14 based on

certain operating parameters. The central station 12 may de-allocate slots for instance where the time since the last use of the slot exceeds some threshold value or if the remote station 14 does not respond to a known number of consecutive MRSYNC frames.

During the next downward period of the data period 51 the central station 12 transmits the local identifier to the remote station 14 through use of a registration clear to send (RegCTS) frame (FIG. 4B). Upon receiving the RegCTS, the remote station retrieves the local identifier and, using the retrieved local identifier, may transmit a Request to Send (RTS) within the designated reserved slot under an implicit polling format during the request period 50 of the next superframe.

Under an alternate embodiment, the remote station 14, upon receipt of a RegCTS may immediately respond by transmitting data. Alternately, a central station 12 may transmit a RegCTS at any time to fill "holes" in the request period (e.g., when a remote station 14 is deactivated or leaves the service coverage area of the network 10).

In general, implicit polling is performed during the request period 50, and explicit polling -- of only those remote stations which requested access to the channel -- is performed during the data period 51.

To initiate the superframe, the central station broadcasts an RSYNC or MRSYNC frame 52 to all the remote stations. The RSYNC or MRSYNC frame is issued periodically, and it defines the start of a number of time slots of the request period. In the preferred embodiment, the central station sends

a RSYNC or MRSYNC frame at least once every second. If there is less data to exchange then the superframe would occur more often, but not more often than once every 100 ms. If there were less data than could be transferred within the 100 ms interval, then the communication channel would be idle for a portion of the 100 ms.

Under an alternate embodiment, an RTS of the remote station 14 specifies the number of data frames it wants to send during the superframe. It is then up to the central station 12 to determine how many times a remote station 14 gets polled. For instance, a central station 12 wouldn't let an entire superframe be "eaten up" by a single station if it requests to be polled too often. Once a request period 50 is complete, the central station 12 has a picture of all upward and downward data periods, and it will divide up the superframe equitably.

A central station 12 may indicate within the RTS frame during the RTS/CTS/DATA/ACK sequence how many frames it will send to the remote station 14 during a superframe. During a DATA/ACK sequence, the use of a "more" bit indicates to the remote station 14 that there is more data to be transmitted during the superframe.

Every remote station has a preassigned waiting period that will begin upon the reception of the RSYNC or MRSYNC frame. These waiting periods are illustrated as time slots TS in Figure 6, which fill up the remainder of the request period 50.

Since remote station 1 has been assigned the first time slot, it issues a reserved slot request RR frame 54 if it has data to transmit on the channel. Hence, the first time slot has

been replaced with the reserved slot request frame RR_1 (RTS frame) transmitted from remote station 1. As seen in the example of Figure 6, no reserved slot request frame was issued in time slot 2 (frame 55), and a reserved slot request frame RR_3 was issued from remote station 3 in time slot 3 (frame 56). In the example shown, a maximum number $X-X_c$ (where X is total slots and X_c is contention slots) denotes the number of active remote stations in the network, and, accordingly, the number of preassigned time slots. (See frame 56.) Note that, in this example, the absence of a reserved slot request frame in a time slot represents a negative access request signal to the central station 12. As will be seen below, an alternate embodiment of the protocol always returns either a positive or negative access request signal to the central station upon issuance of a MRSYNC frame.

After every station has been given a chance to make a reservation, the central station will switch to a modified explicit polling mode, wherein it will sequentially issue a CTS frame to every remote station 14 that made a reservation.

Before the central station 12 begins the explicit polling, on the other hand, the central station 12 must describe the data period 50 for the benefit of those remote stations 14 that may wish to power-down for portions of the data period 50. The central station 12 describes the data period 50 to the remote stations 14 by transmitting a DSYNC or EDSYNC frame 53. (The DSYNC and EDSYNC frames differ primarily in the amount of information provided. In general, the EDSYNC allows for a lower duty cycle of remote stations 14).

If either a DSYNC or an EDSYNC frame 53 is used, then the reader will find via reference to FIG. 4A that the length of the polling period for the upward transmission of data is to be found within the INFO field 34 of the DSYNC or EDSYNC frame 53. A remote station not needing to transfer data to the central station 12 may use the time period specified to deactivate its transmitter and receiver until a point just before the broadcast period, where the remote station 14 must again re-activate its receiver for the receipt of system information during the broadcast period.

As illustrated in Figure 6, the central station polls the first remote station during frame 60 of the upward period with CTS frame P_1, since remote station 1 sent its reserved slot request frame RR_1 during frame 54. Immediately upon receiving the poll signal addressed to remote station 1, that station responds with its data packet $DATA_1$ during frame 62. The central station then checks its poll list to determine which remote station is to be polled next. In the example shown, remote station 3 is polled via poll frame P_3 during frame 64, and it responds with its data packet $DATA_3$ during frame 66. The polling ends upon the completion of the response of the last station on the list, which, in this case, was remote station 3.

Priority message capability is also provided for in the reservation-based polling and data exchange protocol of the present invention. Recall that the control field 32 of the channel frame format 20 (FIG. 2) includes a number of bit-mapped priority flags. In the preferred embodiment, four levels of priority can be implemented using two priority flag bits. If any

remote station had a priority message to send, then that station would set its priority flags to the appropriate priority level, and transmit a reserved slot request RR frame to the central station in its preassigned time slot during the reserved slot request period. Upon receipt of this reserved slot request frame containing priority information, the central station would rearrange the poll list into priority-level order. Accordingly, the central station would poll the remote stations in priority-level order.

The timing cycle diagram shown in Figure 6 can be used to illustrate the reservation-based polling protocol with priority-level polling. Assume that the time slots TS_1, TS_2, TS_3, (frames 54-56) of the reserved slot request period are sequentially assigned to correspond with three remote stations 1-3. If all three remote stations had non-priority messages to send, then each would send its reserved slot request RR frame during the appropriate time slot, and the central station would poll each remote station in numerical order, i.e., the poll list would appear as: P_1, P_2, P_3. If, however, remote station 3 had a level-one priority message to send, and remote station 2 had a level-two priority message to send, then these stations would indicate such using the priority flags in the control fields of their reserved slot request frames. The central station would then re-order its poll list to appear as: P_3, P_1, P_2. Thus, the remote stations are polled in priority-level order. Numerous other multiple-level priority message schemes can be used with the present invention, a few of which will be described below.

Figure 7 represents a similar timing cycle diagram to that of Figure 6, with the addition that an acknowledgement (ACK) frame is transmitted from the central station to the remote station after the reception of each data message from the remote station. In order to send an ACK frame, the central station 12 must, first, correctly receive the data message before issuing an ACK frame (FIG. 4C).

The example of Figure 7 illustrates that, during the reservation request period, remote stations 1 and 3 have transmitted reserved slot request frames 54 and 56, respectively. Therefore, during the upward data transfer period, each of these two remote stations is polled. As before, a first poll frame P_1 is issued from the central station in frame 60, and data packet $DATA_1$ from remote station 1 is returned during frame 62. However, now an acknowledgement frame AK_1 is sent from the central station to remote station 1 during frame 63. A similar polling/data transfer/acknowledgement sequence occurs for remote station 3 during frames 64, 66, and 67. As only partially shown in Figure 7, remote station X-X_n was polled, it transmitted its data packet, and its acknowledgment frame AK_X is shown being returned during frame 69.

If the remote station 14 does not receive an acknowledgement (ACK) from the central station 12 following a data tr[illegible]er (or does not get polled), then the remote station 14 sen[illegible] a reserved slot request (RR) during the next request period 50. If the remote station 14 does not get a response after 3 tries, the data is discarded.

The broadcast period follows the upward period. Any stations which may have de-activated during the upward period must re-activate for the broadcast period. During the broadcast period, data is broadcast from the central station 12 to all remote stations 14. Data frames (FIG. 4C) during the broadcast period are sent with the broadcast DID (e.g., FFFF hexadecimal). Broadcast data frames are not preceded by an RTS/CTS exchange and are not acknowledged by receiving remote stations 14. If there is no broadcast data to be sent from the central station 12 to the remote stations 14, then an EDSYNC frame 53 at the beginning of the data period 51 may be used to indicate a broadcast length of zero.

Following the broadcast period is the downward data period. If the data descriptor 53 at the beginning of the data period 51 were a DSYNC frame, then all remote stations 14 must remain activated during the downward data period.

If, on the other hand, the data descriptor 53 were a EDSYNC frame, then the contents of the EDSYNC would provide advance notification of which remote station(s) 14 would receive data and, therefore, which remote stations 14 would remain activated during the downward data period. Other remote stations 14 not present within the list of the EDSYNC frame may deactivate for the duration of the downward data period.

Data transfer from the central station 12 to the remote stations 14 during the downward period may occur under either of two possible scenarios. The central station may either transmit an RTS and wait for a CTS before transmitting the data, or may simply transmit a data frame and wait for an acknowledgement

response. The use of the RTS by the central station 12 avoids the unnecessary transmission of data when the remote station 14 may not be within range of the central station 12. The use of the RTS/CTS exchange, on the other hand, causes more overall data traffic between the central station 12 and remote station 14.

If the remote station received an erroneous data message, then a negative acknowledgment frame would be returned to the central station. If the central station received neither an acknowledgement frame nor a negative acknowledgement frame from the remote station, then the central station would retransmit the same data message in the next superframe.

Where the RTS/CTS/DATA/ACK sequence is used and there is no response to the RTS, or if the CTS is received with errors, or if after the RTS/CTS/DATA sequence, the ACK isn't received, or if the ACK is received with errors, then the central station 12 begins its retransmission with the retry bit of the RTS frame set. On the other hand, where the DATA/ACK sequence is used and there is no ACK received, or if the ACK is received with errors, then the central station begins its retransmission with the retry bit of the DATA frame set.

Depending upon the requirements of the particular data communication system, it may be advantageous for the central station to track and report on the number of active remote stations in the network -- whether or not each remote station has a data message to send. For this purpose, the central controller would issue a mandatory request synchronization (MRSYNC) frame to all of the remote stations. When a remote station receives this frame, it responds with a RTS frame if it has data to send,

or it responds with a forfeit (FORF) frame if it does not. If a particular remote station does not respond to the global reservation sync frame, then the central station assumes that the particular remote station 14 is not presently active. In this manner, all of the active remote stations will be accounted for by the system without affecting the throughput of the data communications channel.

Periodically, the central station issues a frame (RSYNC, MRSYNC, DSYNC, or EDSYNC) including a superframe number. The superframe number may be used by the remote stations 14 as a functional check of proper operation (e.g., that a particular sleep mode interval did not cause a remote station 14 to miss part of a superframe).

The timing diagrams of Figures 8a and 8b illustrate the sleep mode of remote station 3. During the sleep mode, the controller in the remote station 14 may disable the infrared transmitter and/or receiver circuitry, as well as any other circuitry such as a communications processor which is not being used at the time. This sleep mode ensures minimum power consumption to extend the life of the battery. Figure 8a represents the power consumption of the remote station receiver, and Figure 8b represents the power consumption of the remote station transmitter. These two timing diagrams correspond to the timing cycle shown in Figure 7, wherein acknowledgment frames are utilized.

Since the reservation sync frames 52 and descriptor frames 53 are substantially periodic, the remote station can be programmed to periodically enable its receiver to wait for a

reservation sync frame 52 and descriptor frame 53. Accordingly, as shown in Figure 8a, the receiver of remote station 3 is turned on at time t_0, which precedes the occurrence of the reservation sync frame RS at time t_1 by a sufficient amount to account for clock tolerances. After the reservation sync frame has been received, the receiver is disabled at time t_2.

At time t_3, the transmitter circuitry is enabled such that the reservation request frame RR_3 can be transmitted during time slot 3. At time t_4, the transmitter returns to the sleep mode. At time t_5, the reservation request period has ended, and the polled data transfer period (upward period) has begun.

In general, if the remote station 14 has requested access to the network than at least the receiver needs to remain active during the upward period for the receipt of polling messages. Upon receipt of a polling message, directed to the remote station, the receiver may be deactivated and the transmitter activated. Also, if the descriptor for the data period 51 is a DSYNC frame, then the remote station 14 must remain active for the broadcast period and for the downward period. Further, if a frame directed to the remote station 14 is detected by the remote station 14, then the transmitter of the remote station 14 must be activated for transmission of acknowledgement message.

If the descriptor 53 of the data period 51 is a EDSYNC frame, then the remote station 14 shuts down unless otherwise required. If the remote station 14 has transmitted an access request during the request period, then the receiver of the remote station 14 would remain active until polled, at which time the receiver would deactivate and the transmitter activate for

transmission of the data frame. At the end of the data frame the transmitter would again deactivate and the receiver activate for receipt of the acknowledgement frame from the central controller 12. Likewise, the remote station would only activate for the broadcast period if the EDSYNC message indicated that the broadcast period would have a non-zero time period, or if a data frame were to be directed to the mobile station 14 during the downward period.

Accordingly, remote station 3 (FIG. 8a) must enable its receiver such that it can wait for its poll frame P_3. At time t_6, the poll P_3 has been received such that the receiver can be disabled. However, the transmitter is immediately enabled since data packet $DATA_3$ must be transmitted during frame 66. From times t_7 to t_8, acknowledgement frame AK_3 is being received by remote station 3. After time t_8, the receiver of the remote station can return to its sleep mode until the broadcast period and downward period. Where a DSYNC descriptor 53 is received and if no messages were received by the remote station 3 (as depicted in FIG. 8a) (under either DSYNC or EDSYNC descriptors 53), then at least the transmitter will remain deactivated until the next superframe. As can now be seen, the sleep mode is used by the remote station to conserve battery power when the central station 12 is communicating with other remote stations 14. Various other sleep mode configurations may also be used, particularly since many of the communications processors used in the remote stations may include their own internal power conservation circuits and software.

Referring now to Figure 9, a detailed block diagram of one of the remote stations is shown. As described above, each remote station 14 includes a transceiver which communicates with the central station via an optical-infrared data link 19. The heart of the remote station is a remote controller 110 which, in the preferred embodiment, is a Motorola 68EC000, available from Motorola Corporation, operating at 8 Mhz. Remote controller 110 interfaces with a data processor 112 and a communications processor 114, such that data processor 112 can communicate over the infrared link using the polling protocol described above. In the preferred embodiment, data processor 112 may be part of an EPSON Model No. H1001BEW hand-held computer, and communications processor 114 may be an 82590 LAN interface chip also available from Intel or may be a Field Programmable Gate Array (FPGA) with custom programmed logic provided by Spectrix Corp., of Evanston Illinois.

Communications processor 114, in turn, controls an infrared transmitter 116 and an infrared receiver 118. Infrared transmitters and receivers are well-known in the art. In order to perform the control of the sleep mode for the remote station, remote controller 110 also controls the application of power from power supply 124 to the transmitter and receiver blocks. In the preferred embodiment, power supply 124 is contained within the hand-held computer of the remote station 14. A clock 126 and a memory 128 are also connected to remote controller 110 in order to perform the synchronization and station identification functions of each remote station.

Figure 10 is a detailed block diagram of central station 12 of the data communications network shown in Figure 1. In order to communicate with the remote stations, the central station includes an external transceiver 16. In the preferred embodiment, infrared transceiver 16 is located at a distance from central station 12, since a personal computer is used for the network controller and since the infrared link must be direct line-of-sight. A network controller 130 interfaces an input/output port 132 to a communications processor 134 such that the reservation-based polling protocol of the present invention is used to transmit and receive data from infrared link 19 to I/O port 132 via infrared transmitter 136, infrared receiver 138, and hard-wired link 16. In the preferred embodiment, the function of network controller 130 is performed by an IBM-compatible personal computer using a DOS-based operating system. The personal computer typically includes a memory 140, a clock 142, a display 144, and a keyboard 146.

In review, it can now be seen that the present invention provides an improved contention and noncontention-based multiple-access signalling protocol for a data communications network which efficiently utilizes a single channel even when only a fraction of the users have data messages to send at a given time. The reservation-based polling protocol is particularly adapted for use with a large number of portable battery-powered computer devices communicating with a central station via an infrared link.

While specific embodiments of the present invention have been shown and described herein, further modifications and

improvements may be made by those skilled in the art. All such modifications which retain the basic underlying principles disclosed and claimed herein are within the scope of the invention.

What is claimed is:

1. An apparatus for exchanging data between a central station and a plurality of wireless remote stations on a time divided multiple access communication channel comprising: means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol; and means for polling for data transfers during a second time period the remote stations of the plurality of remote stations providing access requests under non-contention based protocols during the first time period.

2. The apparatus as in claim 1 wherein the means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol further comprises a plurality of time division multiple access slots within the first time interval.

3. The apparatus as in claim 1 wherein the means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol further comprising means, located within a slot of the plurality of slots, for identifying contention slots and non-contention slots to the plurality of remote stations.

4. The apparatus as in claim 3 further comprising means for receiving an access request from a remote station of the plurality of remote stations within an identified contention

slot and allocating a non-contention slot to the requesting remote station.

5. An apparatus for reducing a duty-cycle of activation of a remote station exchanging data with a central station on a wireless time divided multiple access communication channel comprising: means for receiving an access request from the remote station during a first time interval under one of a contention based protocol and a non-contention based protocol; means for exchanging data during a second time period with the remote station providing the access request under the non-contention based protocol during the first time period for data transfers; and means for providing an indicia of epoch length of the first and second time periods to the remote station by the central station.

FIG. 1

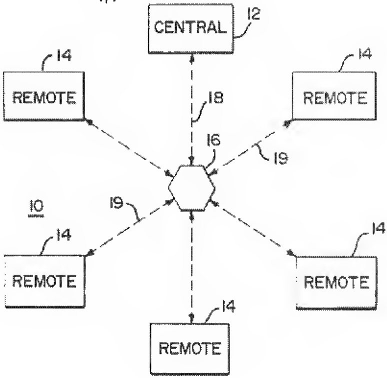

FIG. 2

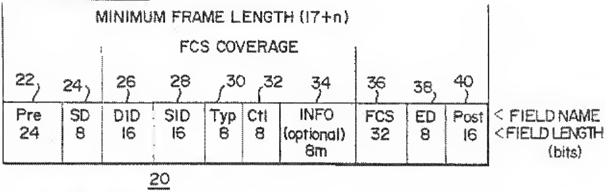

FIG. 3

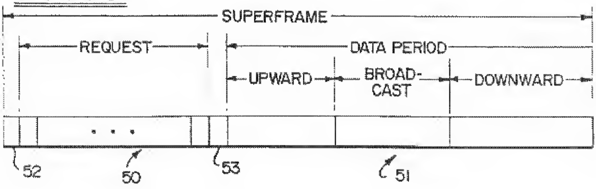

FIG. 5

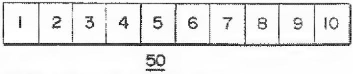

FIG. 4a

RSYNC-Request Sync
MRSYNC-Mandatory Request Sync

	Field	
	Preamble	
	SD	
	DID	< Broadcast
	SID	< ID of originating controller
	Type	< RSYNC or MRSYNC
	Control	< Control flags: AP
	Version	< Software version
INFO	TotalSlots	< Total number of time slots in SyncPeriod (including RegSlots)
	RegSlots	< Total number of time slots which are for registration only
	SuperFrame	< Superframe number
	FCS	
	ED	

DSYNC-Data Sync

	Field	
	Preamble	
	SD	
	DID	< Broadcast
	SID	< ID of originating controller
	Type	< DSYNC
	Control	< Control flags: AP
INFO	UpLength	< length of Upward Data Period
	SuperFrame	< Superframe number
	FCS	
	ED	

EDSYNC-Extended Data Sync

	Field	
	Preamble	
	SD	
	DID	< Broadcast
	SID	< ID of originating controller
	Type	< DSYNC
	Control	< Control flags: AP
	UpLength	< length of Upward Data Period
	BCastLength	< length of Broadcast Data Period
INFO	DownLength	< length of Downward Data Period
	ListLength	< length of following list
	List	< list of stations for which there is downward data in this superframe
	SuperFrame	< Superframe number
	FCS	
	ED	

FIG. 4b

RegRTS-Registration Request

	Field	
	Preamble	
	SD	
	DID	< Destination Station ID
	SID	< registration slot number (temporary station ID)
	Type	< regRTS
	Control	< Control flags: none used
INFO	Version	< Software version
INFO	SA	< Address station registering, 48-bit address
	FCS	
	ED	

RTS-Request To Send

	Field	
	Preamble	
	SD	
	DID	< Destination Station ID
	SID	< Source Station ID
	Type	< RTS
	Control	< Control flags: AP, sequence, out-of-sequence, retry, more
INFO	DataCount	< Number of data frames station wants to send to DA
INFO	DataLength	< Length, in octets, of data the source wants to send
INFO	DA	< Address station to which data is to be sent, 48-bit address
	FCS	
	ED	

FORF-Forfeit

Field	
Preamble	
SD	
DID	< Destination Station ID
SID	< Source Station ID
Type	< FORF
Control	< Control flags: none used
FCS	
ED	

RegCTS-Registration Clear to Send

	Field	
	Preamble	
	SD	
	DID	< Destination Staion ID
	SID	< ID of originating controller
	Type	< ReCTS
	Control	< Control flags: AP
INFO	ID	< ID assigned to station
INFO	SA	< Address station registering, 48-bit address
	FCS	
	ED	

FIG. 4c

CTS-Clear To Send

Field	
Preamble	
SD	
DID	< Desination Station ID
SID	< Source Station ID
Type	< CTS
Control	< Control flags: AP, sequence, out-of-sequence
FCS	
ED	

DATA- Data

	Field	
	Preamble	
	SD	
	DID	< Destination Station ID
	SID	< Source Station ID
	Type	< DATA
	Control	< Control flags: AP, sequence, out-of-sequence, retry, more
INFO	SA	< Address of data originator, 48-bit address
INFO	DataLength	< Length, in octets, of data to be sent
INFO	Data	< Data
	FCS	
	ED	

MDATA-Management Data

	Field	
	Preamble	
	SD	
	DID	< Destination Station ID
	SID	< Source Station ID
	Type	< MDATA
	Control	< Control flags: AP, sequence, out-of-sequence, retry, hierarchical
INFO	SA	< Address of data originator, 48-bit address
INFO	MType	< Type of management message
INFO	Data	< according to MType
	FCS	
	ED	

Acknowledge

Field	
Preamble	
SD	
DID	< Destination Station ID
SID	< Source Station ID
Type	< ACK
Control	< Control flags: AP
FCS	
ED	

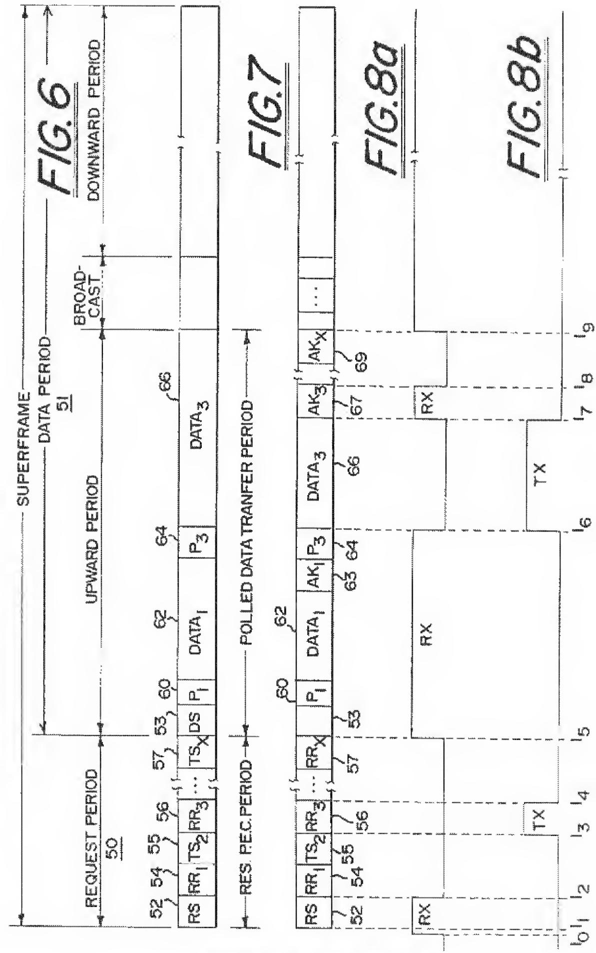

FIG. 6

FIG. 7

FIG. 8a

FIG. 8b

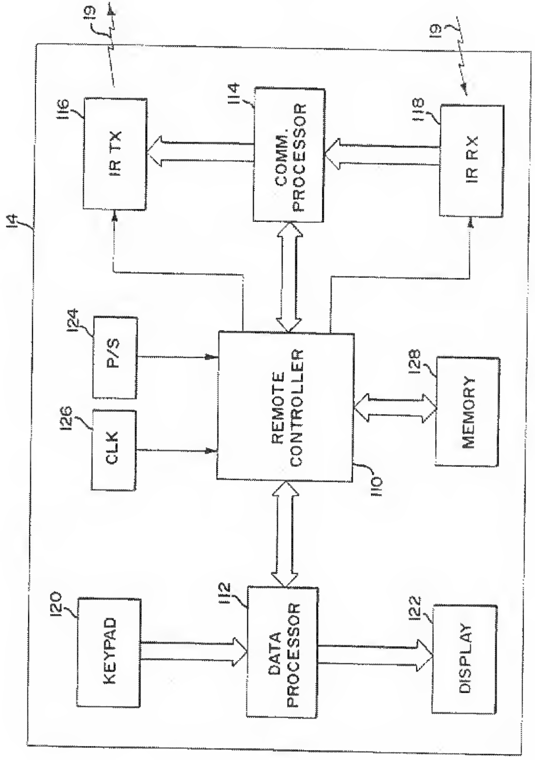
14
19
19
IR TX
116
COMM. PROCESSOR
114
IR RX
118
P/S
124
CLK
126
REMOTE CONTROLLER
110
MEMORY
128
KEYPAD
120
DATA PROCESSOR
112
DISPLAY
122

FIG. 9

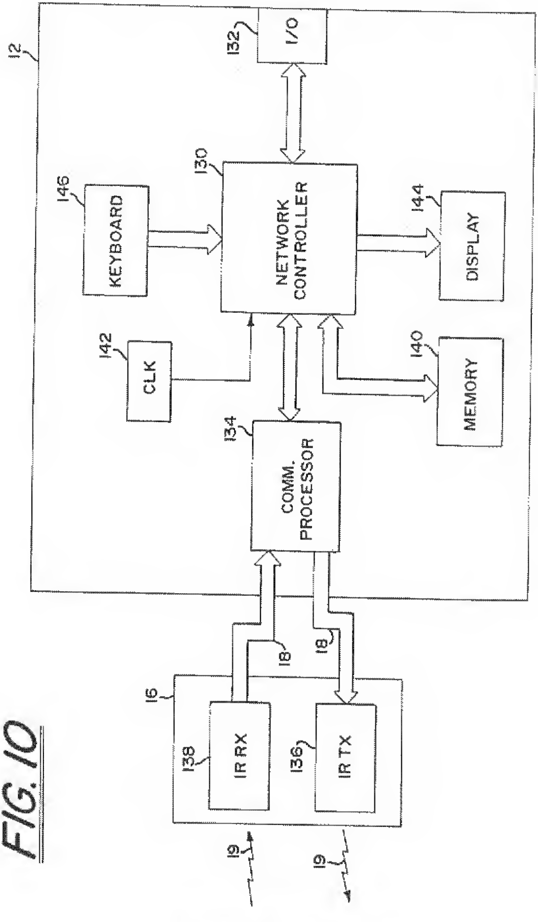
12
I/O
132
130
NETWORK CONTROLLER
146
KEYBOARD
144
DISPLAY
142
CLK
140
MEMORY
134
COMM. PROCESSOR
18
18
16
IR RX
138
IR TX
136
19
19

FIG. 10

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US95/05612

A. CLASSIFICATION OF SUBJECT MATTER
IPC(6) : H04J 3/16
US CL : 370/095.200 , 095.300 , 085.800 , 085.200 ; 340/825.080
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 370/095.200 , 095.300 , 085.800 , 085.200 , 085.700 , 095.100 ; 395/200.090 , 455/004.100; 340/825.080

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
NONE

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
NONE

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US, A, 5,297,144 (GILBERT ET AL.) 22 March 1994, columns 1-6 and figures 1-4.	1-5

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

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Date of the actual completion of the international search	Date of mailing of the international search report
05 OCTOBER 1995	03 NOV 1995

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